Lucas

TOPIC: VOLTAGE CONTROL OSCILLATOR

The transistorized Voltage Control Oscillator was developed to combine ease of set-up and calibration with a maximum versatility. The VCO-3A is a blocking oscillator whose output frequency can be varied as a function of the amplitude and polarity of some control voltage. It thus operates equally well with either ac or dc input signals, giving it a wide range of applications. It can be used to directly monitor operating or supply voltages, or it can be used with any of a wide variety of transducers whose outputs are ac or dc signals proportional to the physical quantity being measured. The VCO-3A oscillates over a wide range of frequencies and is used on channels 3 through 18. It is used with high frequency pickups such as vibration transducers.

Set-up and calibration of the oscillator are exceptionally easy. The sensitivity of the oscillator is adjusted for 88% of bandwidth with standard voltage inputs. Any transducer can then be used with the VCO-3A, providing that the excitation voltage to the transducer has been adjusted so its maximum output is equal to the standard calibration input. As the calibration of the oscillator and the transducer are independent of each other, either component can be replaced in the field without requiring recalibration of the other.

The operating frequency of the oscillator is independent of transistor characteristics. Careful circuit design, along with temperature compensation of the RC networks, has achieved stable performance even in the critical POLARIS environment.

THEORY OF OPERATION

Figure 10 is a simplified block diagram of the VCO. The frequency control circuitry consists of a loop which is formed by the blocking oscillator, the transistor switch Q1, the RC network, the pick-off diode CR1, and the emitter-follower trigger-amplifier combination. In the normal course of its operation the blocking oscillator, which is free-running, sends a positive trigger pulse to the transistor switch Q1. This closes the switch and shorts condenser C4-C20. The voltage at the junction of R8-R9 and C4-C20 quickly falls to a negative value, cutting off CR1.

After the positive pulse from the blocking oscillator ends, the transistor switch Q1 cuts off again and the condenser C4-C20 starts charging toward a positive voltage. As soon as the voltage at the junction of R8-R9 and C4-C20 rises to a few millivolts above the input voltage, CR1 conducts and couples the rising voltage to the emitter-follower trigger-amplifier combination. Here it is amplified and inverted, appearing in the output as a large negative going signal. This signal is coupled to the sync input of the blocking oscillator. The free-running period of the blocking oscillator is 20% to 50% longer than the charge time of condenser C4-C20, so the arrival of the trigger pulse causes it to fire early. The triggering of the blocking oscillator initiates another pulse which goes to the transistor switch Q1. This abruptly terminates the charging of condenser C4-C20 and thus starts another cycle.

The period between these cycles, is of course, a function both of the RC time constant of the frequency control network, and of the voltage at which CR1 starts to conduct. The output of the VCO-3A can thus be frequency modulated by applying a control voltage (0, ± .75 volts) which will vary the bias on CR1. Figure 13 shows how the application of either a negative or positive signal to the input connection changes the period of oscillations. As the center-frequency of the VCO-3A is controlled by the RC network, shown in the simplified block diagram, and not

by the free-running blocking oscillator, the frequency is thus independent of the characteristics of the transistors in the blocking oscillator circuit - hence no selection of transistors is required. The frequency will vary, of course, with supply voltage variations or with temperature. The two Zener diodes, CR3 and CR4, control the charging voltage, EB, and make it independent of supply voltage. Temperature compensation is used to eliminate RC drift with temperature.

The output of the blocking oscillator, is a saw-tooth voltage which is coupled to a clipper-amplifier. The output of this clipper-amplifier is essentially a square wave, having very little even harmonic content. By eliminating the second harmonic, a more compact band-pass filter can be used.

DETAILED CIRCUIT ANALYSIS

Figure 11 is a partial schematic of the frequency control portion of the VCO-3A, including everything in the basic loop but the blocking oscillator. The two Zener diodes hold the voltage across the RC network to approximately 29 volts. When C4 is completely discharged, the voltage at the junction of R9 and C4 is -22 volts. C4 then charges towards a positive voltage whose maximum amplitude, (as set by R4), is 7 volts. C2 and C8 are filter condensers, preventing surges from either C4 or the blocking oscillator from getting into the regulated voltage.

During the initial set-up of the oscillator, R4 is adjusted for 7 volts. The center frequency is then set by selecting C4, C20, R8 and R9 for the channel desired. After channel calibration is completed, R4 can then be used to adjust the frequency range for use with a certain type of transducer. If the output of a transducer, for example, is ± 0.75 volts, then the R4 setting would not be changed. However, if the output of the transducer were 0 to + 1.5 volts, then R4 would be adjusted so that center frequency was obtained with an input signal of +0.75 volts. Figure 14 shows how R4 changes the frequency by changing the charging voltage.

R8 and R9 are a carbon film and a wirewound resistor in that order, and are used for temperature compensation of the frequency control network. They also compensate for temperature effects in the two Zener diodes, both CR3 and CR4 having positive temperature coefficients. R7 provides a FINE adjustment on frequency, giving a maximum change of approximately 10% of band-width.

Since the base of Q1, the transistor switch, is at the same potential as its emitter, it is normally "off". When the 5 volt pulse arrives from the winding of transformer (T1), Q1 is turned "on" and shorts C4, C20. Base current develops a negative bias across R11 and C5, which sharply cuts off Q1 upon removal of the positive 5 volt pulse. R10 is a current limiting resistor, added to protect the transistor.

The cathode of the pick-off diode, CR1, is at the input signal potential. This diode is "off" until the voltage across C4 rises to a value slightly positive in respect to the input signal potential. CR1 then conducts and couples the rising voltage, through C6, to the base of the emitter-follower Q2. The voltage across C4 continues to rise until it is between 20 millivolts and 40 millivolts positive in respect to CR1 cathode, at which time triggering of the blocking oscillator occurs and C4 is discharged. The large negative spike following this small positive rise is due to differentiation action by the internal capacity of CR1. The discharge of C4 is coupled through this capacity (CR1 internal) to the base of the emitter-follower shown on the waveshape as a sharp negative spike. R13 provides a sensitivity adjustment. During initial set-up a standard calibration

voltage of + 0.75 volts is applied to the input and R13 is adjusted to give a frequency deviation of + 44% of bandwidth. The Thermistor provides temperature compensation for sensitivity. Figure 15 represents a plot of output frequency versus temperature for three different calibration voltages. The change in frequency due to the temperature sensitivity of the circuit can be compensated for by raising the input voltage at higher temperatures. The Thermistor, having a negative temperature coefficient, allows an increasing portion of the input signal to be applied to CR1 as the temperature rises.

Q2 is an emitter-follower and provides isolation between the trigger-amplifier and the frequency control network. It has unity gain and its output signal, developed across R16, is coupled through C7 to the base of Q3. Q3 has a high gain and injects a large negative going spike into the collector circuit of the blocking oscillator Q4. This injection of the triggering pulse is shown in detail in the partial schematic of Figure 12. The time at which triggering occurs is a function of the rise-time of the negative spike. Therefore it is desireable that Q3 have as high a gain factor as practical.

The trigger pulse is coupled, as a positive pulse, through T1 to the base of Q4, causing it to conduct. The voltage developed by current flow through R21, C10 and C11 in the emitter curcuit, however, is large enough to cut-off Q4 at the end of the positive trigger pulse. The large pulse of current, which charges C10 and C11, biasing off Q4, also excites the tank circuit (T1) into oscillation. The second half-cycle of oscillation is absorbed by CR2 and Q4 remains off. After cut-off, the voltage across C10 and C11 drops exponentially from its peak value until the next trigger arrives, when the condensers charge again. Thus a 3 volt peak to peak saw-tooth voltage is developed which is coupled through C13 to the squaring amplifier Q5.

Q5 is biased at near saturation current and base current flows during the positive portion of the saw-tooth signal. Double clipping thus occurs, the base current cutting off the positive swing, and the negative swing cuts off Q5. The result is approximately a square-wave signal in the collector circuit, having an amplitude of 18 volts peak to peak. By squaring the wave-form, the even harmonic content is considerably reduced, enabling more efficient filtering with a smaller, more compact band-pass filter. R28 provides an output adjustment.

The amplitude tell-tale permits monitoring of some event (such as stage separation). The connection is normally shorted by a break-wire. By grounding one end of R31 the negative supply voltage is reduced by the divider formed by R5 and R31. Upon opening the break-wire, a greater supply voltage is applied across Q5, increasing the output signal 30% ± 5%. This amplitude modulation of the output thus provides a record of event time.

LOCKHEED AIRCRAFT CORPORATION
MISSILE SYSTEMS DIVISION

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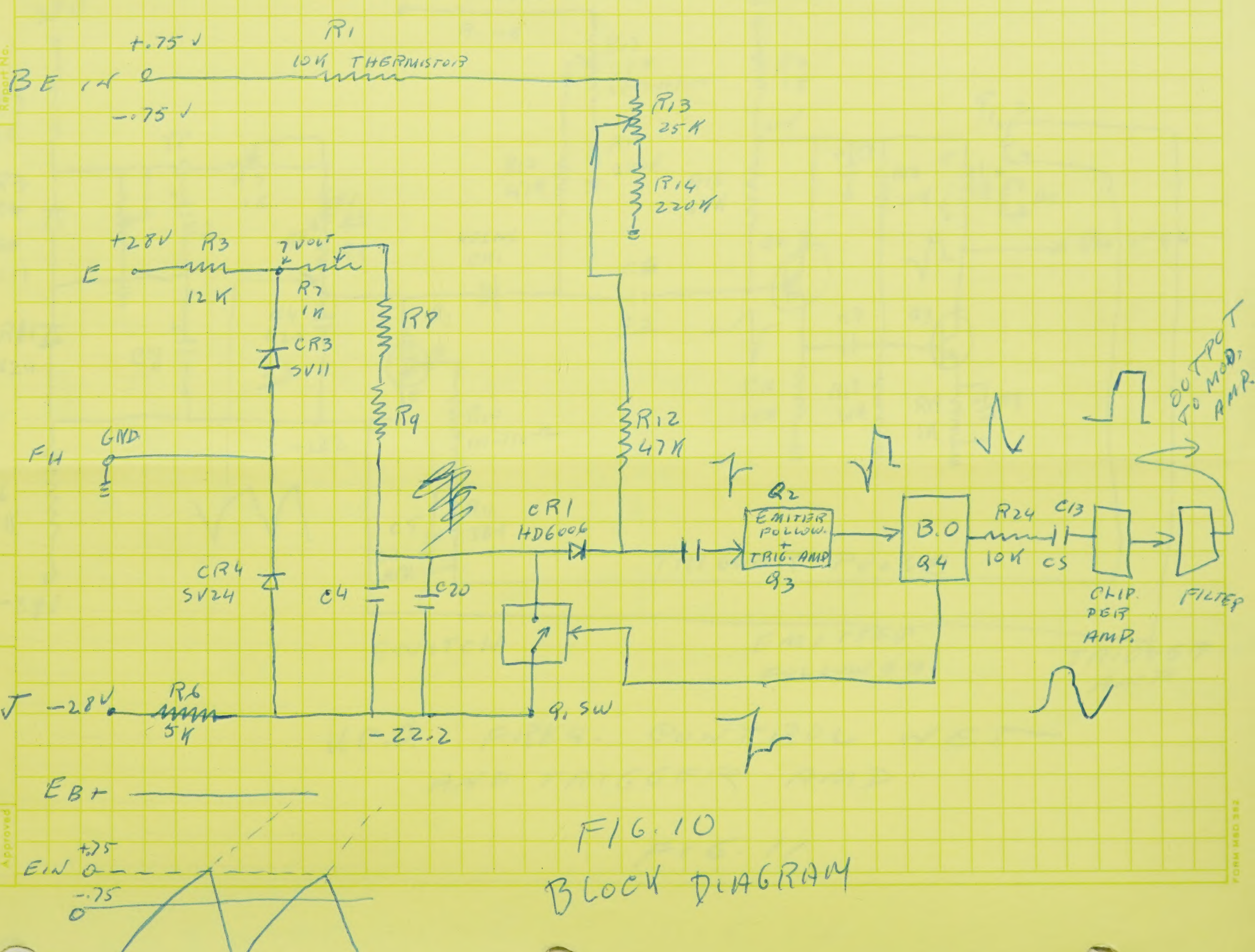

FIG. 10
BLOCK DIAGRAM

FORM MSD 382

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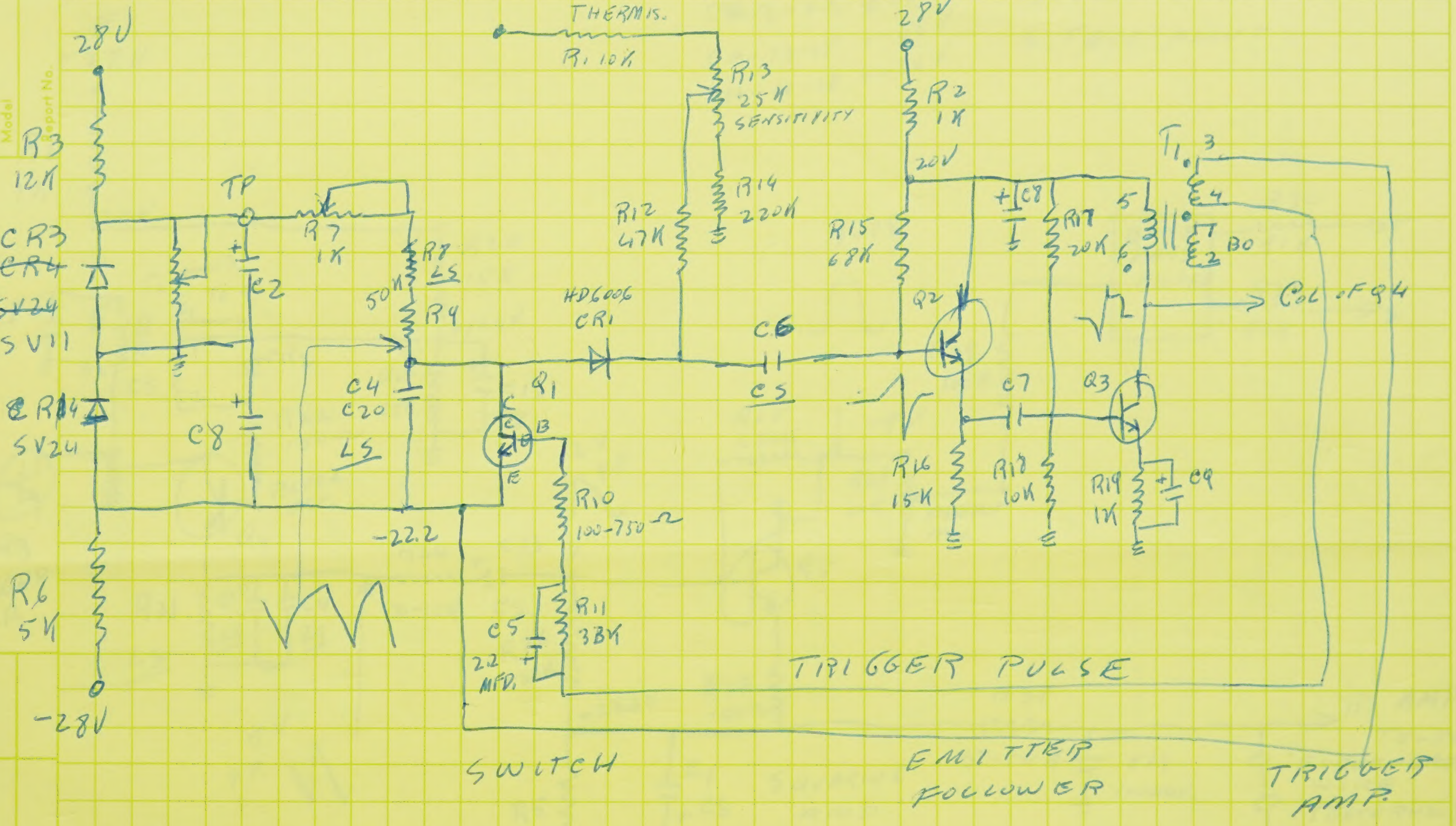
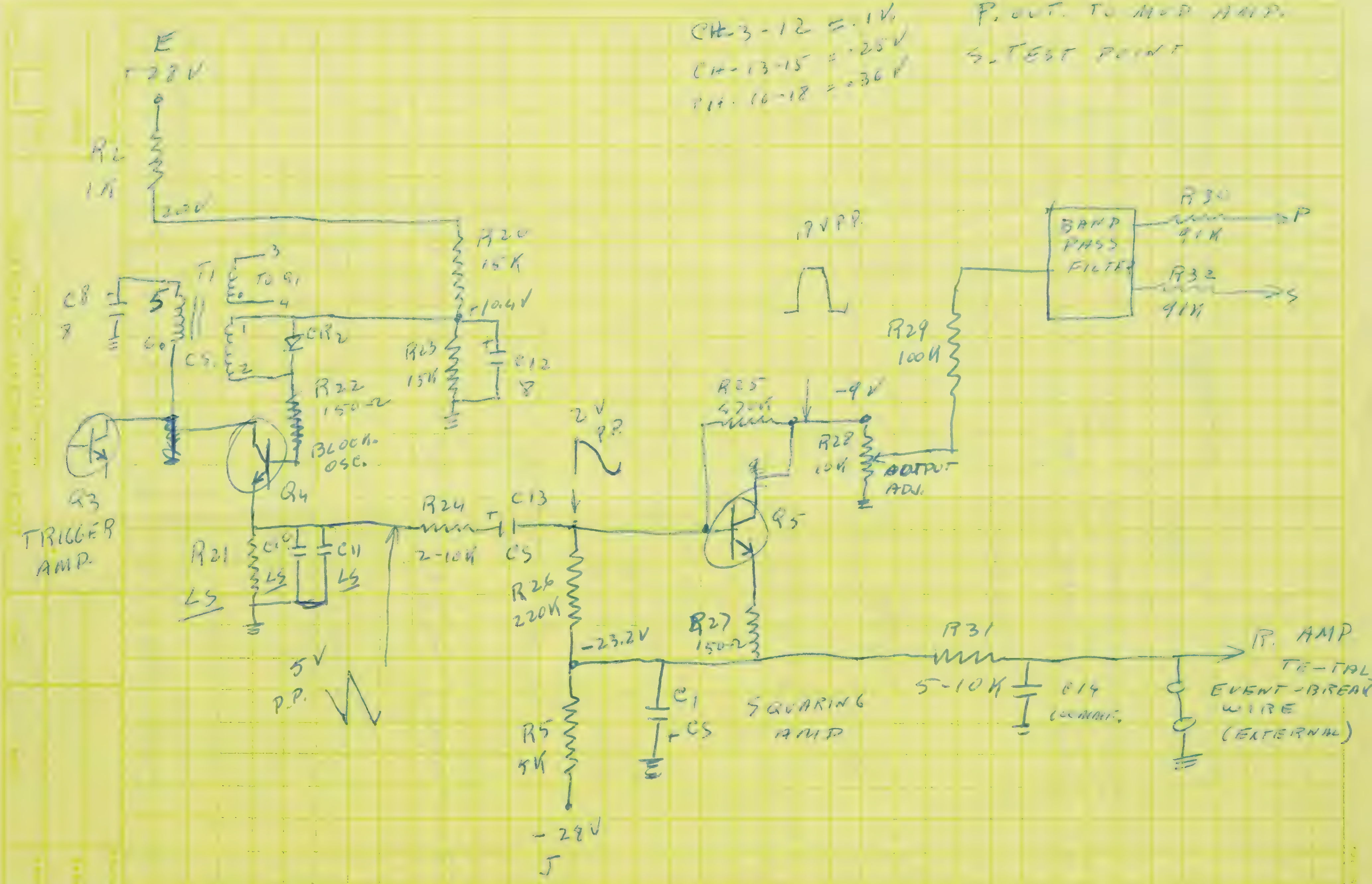


V.C.O. FREQ. CONTROL NET
AND TRIGGER AMP

FIG. 11

FORM MSD 362



BLOCKING OSC. / SQUARING AMP
& OUTPUT FILTER.

FIG. 12

TEST SPECIFICATION F-13
TRANSISTOR Q_1 FOR VCO.

1. This specification describes the minimum standards for Q_1 transistors of VCO.

2. This specification is for the selection of silicon transistors to meet special characteristics for temp. compensation of VCO-3A.

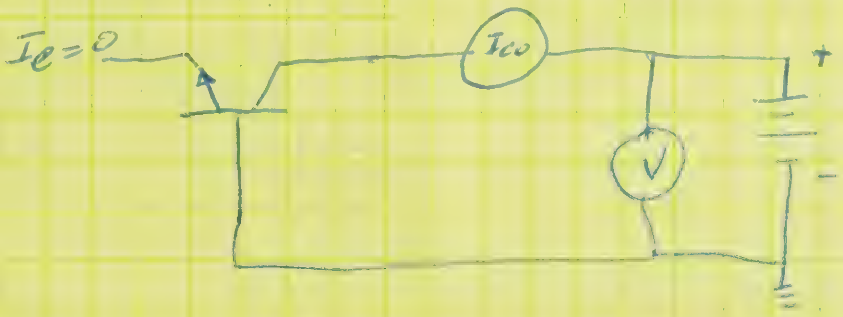
3 FOR NPN silicon Transistors

Tex. INST. 2N 332, 2N338 FAIRCHILD 2N696 OR Equivalent.

4. Test Panel

4.1 Transistor Test Panel
D.C. Microampmeter
VTVM
P.S. 30V DC

5. DIAGRAM.

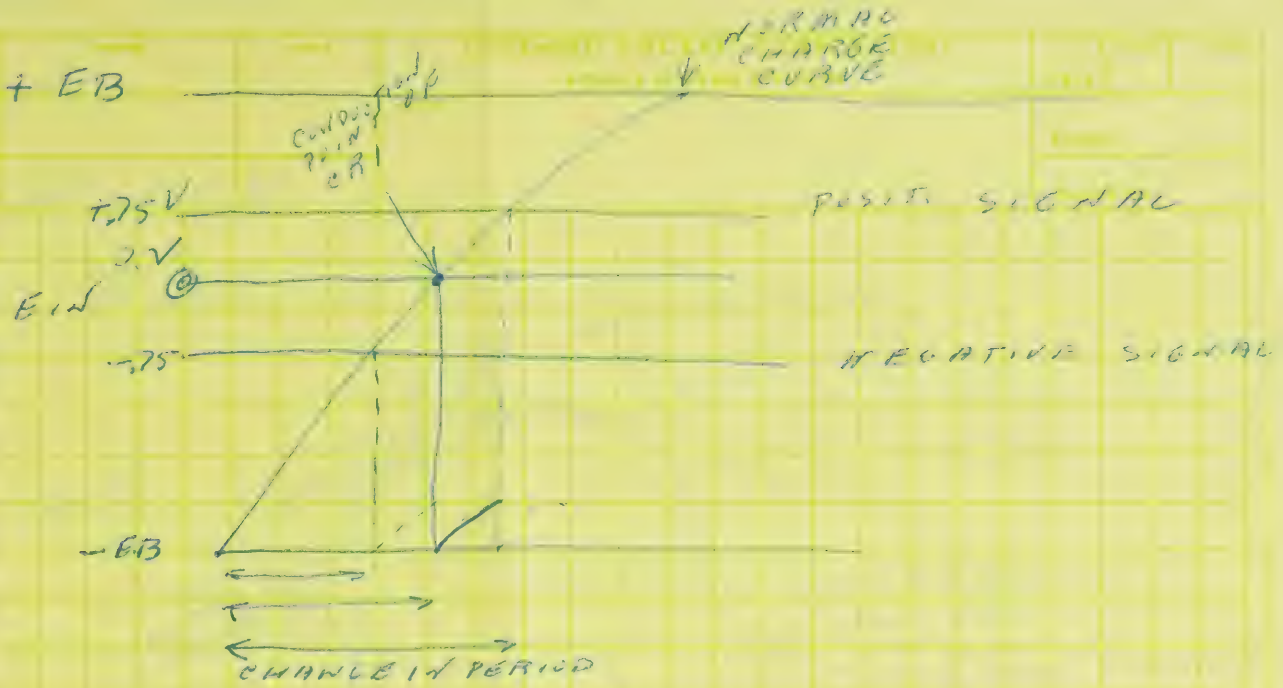


6. Each Transistor shall be tested for conformance to this specification

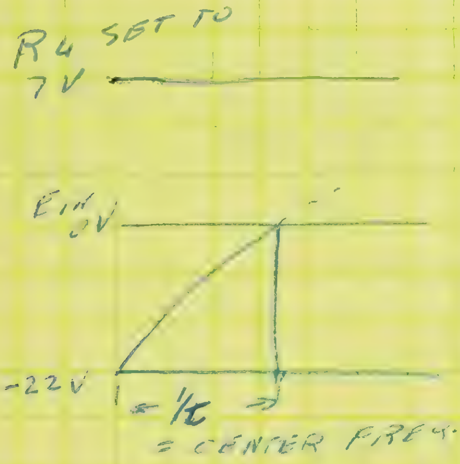
7. Transistor for Q_1 of VCO shall be selected for suitable characteristics of I_{co}

8. with the transistor connected as shown in the diagram raise the temp to 85°C and apply test voltage of +30 V DC

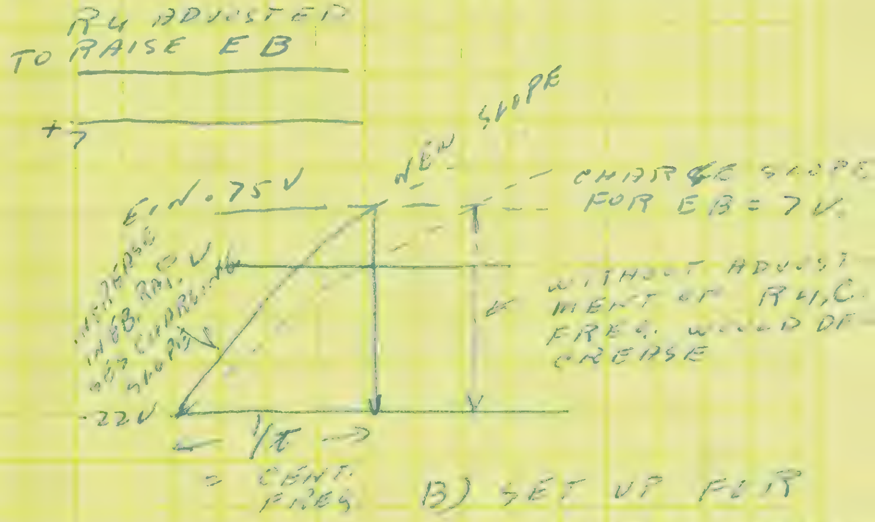
9 The leakage current I_{co} should be less than .7 ma $I_e = 0$ (transistors with leakage currents higher may be used for other parts in VCO-3A)



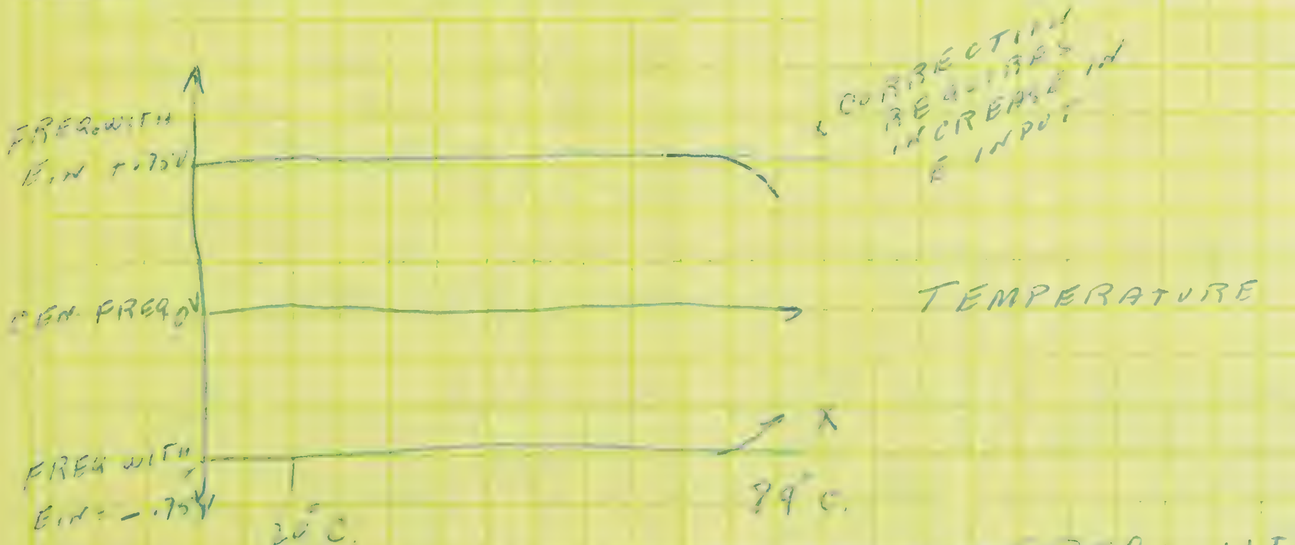
FREQ. MOD. OF V.C.O.



A) SET UP FOR ±.75V TRANSDUCER CENTER FREQ SET WITH EIN = 0V.


R4 ADJUSTED TO RAISE EB
+7
NEW SLOPE
EIN .75V
CHARGE SLOPE FOR EB = 7V.
INCREASE IN EB RAI. 0V CURRENT AND SLOPE
WITHOUT ADJUSTMENT OF R4, C. FREQ. WOULD DECREASE
-22V
1/t
= CENT. FREQ.

B) SET UP FOR 0 TO +1.5 V. TRANSDUCER CENTER FREQ. WITH EIN = +.75V

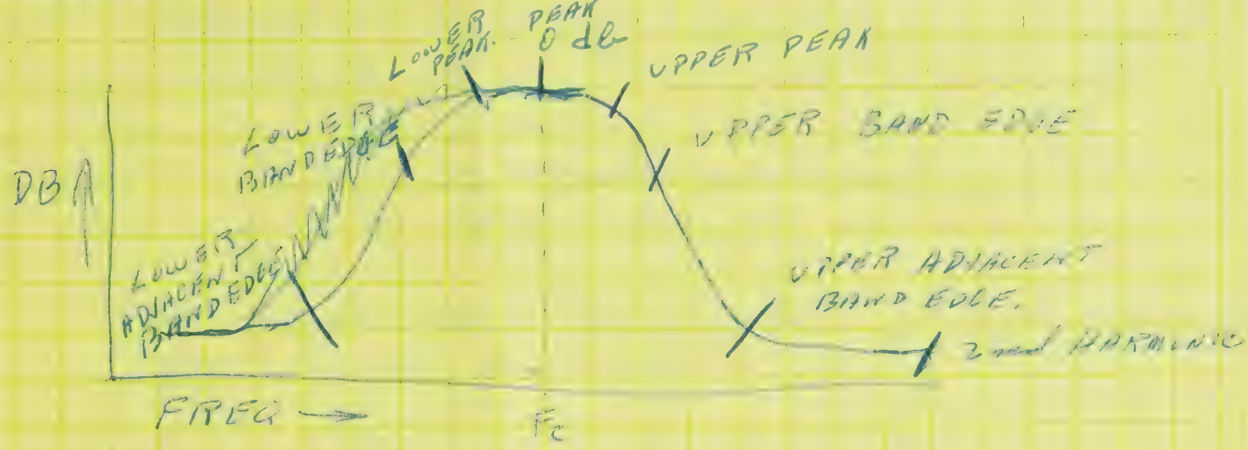


CHANGE OF FREQ. WITH TEMP.

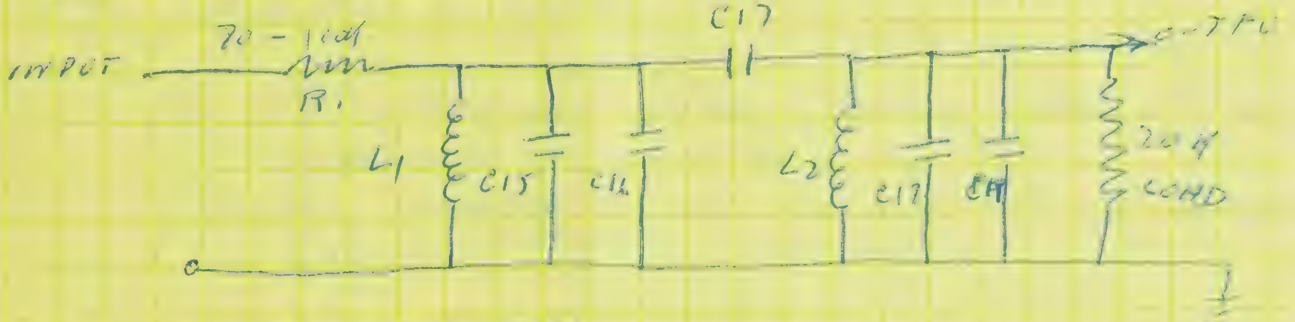
F16-13

FILTER

1 BASIC FILTER CURVE



2 BASIC CKT



C1 = C15 - C16

C2 = C17

C3 = C18 - C19

3 MID BAND RESPONSE

a IF the mid band response is less than the lower or upper band edges ~~decrease~~ C-17 ↓

b If the mid band response is barely detectable, increase C17 ↑

c To shift the mid band freq response to the right or to increase the freq decrease C15 - C16 and C18 - C19 by equal increments in small amounts

5 BAND PASS

a As the Band Pass increases (To WIDE)
C17 ↑

b As the Band Pass decreases (To NARROW)
C17 ↓

C15

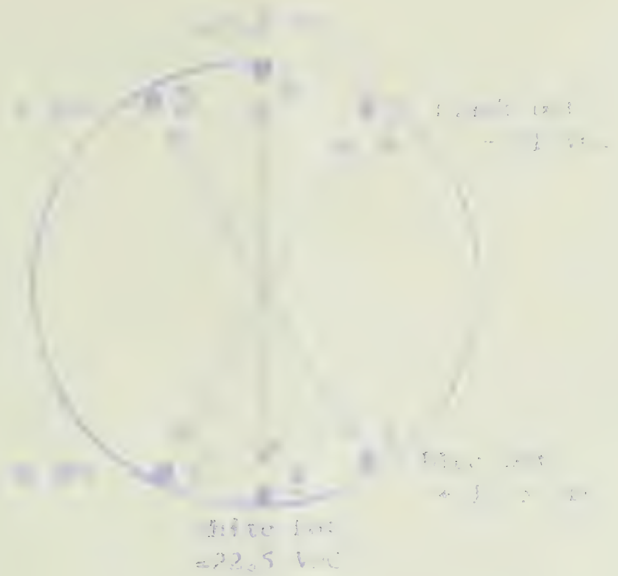
To decrease Band Pass ↓ B.P C17 ↑
To increase Band Pass ↑ B.P. C17 ↓

The minimum band response upper band edges and the lower band edges should be approx. equal.

c Clip. in small value D.M. caps into the correct portion of the filter repeat and zero the HP. LOAD and then make normal filter checks.

[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
			[illegible]	[illegible]		[illegible]	[illegible]	[illegible]	1	[illegible]	.01 µH	3
1	[illegible]	[illegible]	241	0.27 "		15 K	0.22 "	2	1	1.5 "	.01 "	4
	[illegible]	1000	2100	[illegible]	"	20 K	[illegible]	5	1	1.5 "	.01 "	5
	[illegible]	[illegible]	[illegible]	[illegible]		[illegible]	[illegible]	[illegible]	1	[illegible]	.01 "	6
	[illegible]	[illegible]	[illegible]	[illegible]	"	15 K	[illegible]	[illegible]	1	[illegible]	.01 "	7
[illegible]	[illegible]	[illegible]	2150	0.0327"	"	22 K	0.0497"	[illegible]	1	[illegible]	[illegible]	8
		[illegible]	[illegible]	[illegible]		[illegible]	[illegible]	[illegible]	1		[illegible]	9
			[illegible]	[illegible]		[illegible]	[illegible]	[illegible]	[illegible]		[illegible]	10
[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	"	[illegible]	[illegible]	[illegible]	1	0.22 "	1200 µµf	11
[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	"	22 K	[illegible]	[illegible]	[illegible]	0.22 "	1200 "	12
		[illegible]	[illegible]	[illegible]		[illegible]	[illegible]	[illegible]	[illegible]	[illegible]		13
[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	"	20 K	[illegible]	[illegible]	1	0.1 "	[illegible]	14
	[illegible]	[illegible]	[illegible]	[illegible]		[illegible]			[illegible]	[illegible]	[illegible]	15
[illegible]	[illegible]	[illegible]	[illegible]	[illegible]								16
												17
			[illegible]			[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	18
												[illegible]

[illegible]



These pulse transformers are quite fragile and expensive. If a lead is pulled from the outer seal of the transformer - by pulling a lead to one side - the transformer generally becomes defective.

Checking The Transformer Before Installation

The various transformers used on the VCO can be and should be checked prior to installation on the board.

Use either a VOM or a VTVM on the times one [illegible] as follows:

A - Across Pins 1 and 2.

B - Across Pins 3 and 4.

C - Across Pins 5 and 6.

Note: You measure across opposite leads on the transformer.

The following values were taken from an average of [illegible] transformers in each batch series. A new pulse transformer [illegible] readings which are very close to those in the following [illegible]. [illegible] of values for the transformers, which were [illegible] (from 0.1 to [illegible]).

[illegible]

[illegible]

[illegible] [illegible]

[illegible]		[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
[illegible] 2	————	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
[illegible] 4	————	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
[illegible] 8	————	21.0	11.5	[illegible]	[illegible]	2.

[illegible] mounting on the [illegible] a [illegible] [illegible] by [illegible] resistance in the same manner [illegible]

[illegible]: [illegible] the transformer [illegible] versed [illegible] [illegible]

[illegible] for the various [illegible] on page [illegible] [illegible]

LUCAS

FM-SUB-CARRIER CHANNEL TOLERANCES

Revised 8-16-5[illegible]

Set-Up and Check Out

	- DATA LIMIT			- CALIBRATE			MIDBAND			+ CALIBRATE			+ DATA LIMIT			
	-Tolerance -5% BW	-50% Input -42.5%BW	+Tolerance +5% BW	-Tolerance -2.5%BW	-Cal -30% BW	+Tolerance -2.5%BW	-Tolerance -3% BW	Cal Input	+Tolerance 3% BW	-Tolerance +2.5%BW	+Cal +30% BW	+Tolerance +2.5%BW	-Tolerance -5% BW	+50% Input +42.5%BW	+Tolerance +5% BW	[illegible]
	371.5	374.3	377.5	377.5	379.0	380.5	396.5	400	401.8	419.5	421.0	422.5	422.5	425.5	428.5	60
	520.1	524.3	528.5	528.5	530.6	532.7	551.5	560	563.5	587.3	589.4	591.5	591.5	595.7	599.9	[illegible]
	677.7	683.2	686.7	688.8	690.5	694.3	725.7	730	733.3	768.8	768.5	771.2	773.3	775.8	782.3	110
	[illegible]	[illegible]	906.0	906.0	909.6	913.2	955.7	960	964.3	1006.8	1010.4	1014.0	1013.9	1021.2	1028.3	[illegible]
	[illegible]	[illegible]	1227	1227	1232	1238	1294	1300	1306	1364	1369	1374	1373.	1383	1393	195
6	1478	1590	1604	1604	1610	1617	1692	1700	1708	[illegible]	[illegible]	[illegible]	1795	1809	1822	[illegible]
7	2136	2153	2170	2170	2179	2188	2290	2300	2310	2412	2421	2430	2430	2447	2464	345
	[illegible]	2809	2832	2831	2843	2854	2987	3000	3014	[illegible]	3159	[illegible]	3169	3191	3213	450
	[illegible]	3651	3686	3690	3696	3714	3882	3900	3918	[illegible]	[illegible]	[illegible]	4120	[illegible]	[illegible]	[illegible]
	[illegible]	[illegible]	5096	5096	5117	5137	5376	5400	5424	5663	5684	5704	5704	5744	5784	810
	[illegible]	[illegible]	6937	6937	6964	6992	7317	7350	7383	7700	7736	7763	7763	7818	7873	[illegible]
	[illegible]	9810	9899	[illegible]	[illegible]	[illegible]	10453	10500	10547	[illegible]	[illegible]	[illegible]	11090	11170	11249	1575
	[illegible]	[illegible]	13684	[illegible]	[illegible]	[illegible]	14435	14500	14565	15207	15262	15316	15326	15429	15534	[illegible]
	[illegible]	[illegible]	20763	[illegible]	20845	[illegible]	21501	22000	22099	23072	23155	[illegible]	23237	23402	23567	3300
	[illegible]	[illegible]	19525				21670	22000	22330				24475	24805	25130	[illegible]
	[illegible]	[illegible]	28311	28311	28425	28538	29865	30000	30135	31462	31575	[illegible]	31687	31912	32137	[illegible]
	[illegible]	[illegible]	25691				29730	30000	30270				34241	34500	34759	[illegible]
	[illegible]	[illegible]	37750	37750	37900	[illegible]	39820	40000	40180	[illegible]	42100	[illegible]	42250	42550	42850	[illegible]
	[illegible]	[illegible]	35085				39400	40000	40600				44512	45112	45712	[illegible]
	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	52264	52500	52736	[illegible]	[illegible]	[illegible]	55453	55687	56281	[illegible]
	[illegible]275	[illegible]620	[illegible]				52027	52500	52973				59927	60380	60833	[illegible]
	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	69685	70000	70315	[illegible]	[illegible]	[illegible]	73937	74462	74987	[illegible]
	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	69370	70000	70630	[illegible]	[illegible]	[illegible]	77875	78925	79975	[illegible]
							NORMALIZED CHANNEL DIGIT									
	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	20090	20000	19910	[illegible]	[illegible]	[illegible]	18951	18801	18651	
							CHANNEL E DIGIT									
	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	20000	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	

Lettered channels (A,B,C,D,E) are wide band channels BW= 30% f_c

Numbered channels (1 thru 18) are standard band channel BW= 15% f_c

	NAME	DATE	**SPECIFICATION**	LOCKHEED AIRCRAFT CORPORATION MISSILE SYSTEMS DIVISION VAN NUYS, CALIF.
Prepared			TYPE	
Approved			TITLE	
Approved				
Approved				SHEET OF

REVISED:

FORM MSD 875

	ATT.	FIXED VALUES (+)		[illegible]	C. [illegible]	(= 1.5 DB.		)= 10 DB, Ch 6-E		)= 25 DB.		
	C15 & C16	L1 & L2	(C)	C13 & C14	CF	LSB	USB	LSB	USB	2nd. H.	3rd. H.	CH
	.0646 uf	10	.0062 uf	.0144 uf	730	[illegible]	765	640	550	[illegible]	[illegible]	3
	.0159 "	330mh	.0064 "	.0032 "	460	336	1047	785	1200	2520	[illegible]	4
	.016 "	150 "	.0034 "	.019 "	1300	1086	1780	1312	1570	2660	[illegible]	5
	.01009"	190 "	.0018 "	.01075"	1700	1572	2328	2125	2017	3400	[illegible]	6
	[illegible]	[illegible]	.0011 "	.01062 "	2000	2177	2473	1880	2775	4400	6000	7
	.0057 "	140 "	.001 "	.0057 "	3000	2775	3385	2613	3507	6000	9000	8
	.0038 "	130 "	605 uuf	.0036 "	4600	3607	4468	3235	4692	7800	11700	9
	.0017 "	300 "	130 "	.0021 "	5000	4795	5800	3775	6100	10060	16000	10
	1150 uuf	200 "	836 "	1361 uuf	7000	6095	7801	5800	9742	15600	23850	11
	[illegible]	[illegible]	146 "	812 "	10500	9712	11586	7900	13611	21000	31500	12
	[illegible]	150 "	116 "	668 "	14500	13412	15586	11188	18355	29000	43500	13
	[illegible]	100 "	86 "	435 "	20000	20350	23650	15500	27750	40000	60000	14
	[illegible]	70 "	56 "	308 "	30000	27750	32250	25650	37000	60000	90000	15
6	205 "	40 "	40 "	205 "	40000	37000	43000	32250	49920	80000	120000	16
[illegible]	155 "	41 "	32 "	155 "	50000	48500	56450	43000	64750	100000	150000	17
[illegible]	125 "	50 "	26 "	125 "	70000	61750	73250	52450	80000	140000	210000	18
	105 "	30 "	— —	114 "	75000	69500	80500	62250	97000	160000	310000	E

— [illegible] SPACE FOR FURTHER INFORMATION

E. W. [illegible]
1-30-69

FILTER DATA NOTES

⊕ Values are fixed for the coils and coupling condenser and should not be changed, except as a last resort in rework. Average values for C 15, C 16, C 18, and C 19 may be adjusted for correct attenuation in rework. Changes should be made in very small increments, especially on high channels.

✳ This is the approximate frequency to which the H-P 400-D is peaked. (Note: This mid-band frequency may deviate by as much as 3000 cycles from these values, as in the case of CH. #18. In any event be certain you tune for the highest peak - if more than one peak is present.) After peaking, the DB. meter needle is set on 0.0 DB. by adjusting the amplitude knob. Record this frequency as read on the counter, which should be pluged into the H-P 200 - CD oscillator. This becomes the mid-band frequency and all attenuation values, for L&D channel specification are read in reference from this point.

J. W. Simpson
1-30-59

Burch

TELEMETER RADIO FREQUENCY SUB SYSTEM, EXCLUSIVE OF ANTENNA

I. Modulation bus.

Since the over-all signal-to-noise ratio possessed by the telemetering sub-carriers when injected into the transmitters is of paramount importance, as is the fidelity of each subcarrier itself, the R.F. Sub-system might well be considered as beginning at each signal source, or at the output of each channel band-pass filter.

Between each filter and the transmitter itself there exists one or more isolation resistors, one or more common information lines called "modulation buses", and one or more summing amplifiers providing at least one transmitter with subcarrier information.

In practice, for the sake of redundancy related to highly important channels of information, the same sub-carrier filter may furnish the same isolated output to two distinct modulation buses providing signal to two separate transmitters. In the case of the X-17 vehicle, however, each subcarrier furnished its output to two distinct modulation buses which terminated in separate summing amplifiers. The latter, in turn, furnished in-phase, isolated outputs to a single transmitter. This last scheme was utilized as one measure to sharply reduce the possibility of losing information conveyed on long-line modulation buses when the lines themselves were constrained to pass through many plugs, to be subjected to the ejection of vehicle stages and to the severe re-entry environment.

The modulation bus itself should inherently be a noise-free line, since its signal is of relatively low-level. Its long length and its proximity to transient and noise producing areas make its signal-to-noise ratio quite important. Adequate shielding minimizes the external signal pickup, and the very low impedance of the summing amplifier further sharply reduces the effect of stray currents induced in the line.

II. Sub-carrier Summing Network.

Any summing network must be extremely linear in operation to prevent distortion and must furthermore present an impedance ratio conducive to a fidelity summing operation.

The Lockheed summing amplifiers, both vacuum tube and transistorized versions, possess these basic characteristics besides other advantages. By utilizing a large feedback factor, they are extremely/(flat frequency response essentially from 20 CPS to 100 KC., possess a maximum total linear distortion figure of 0.5% and present a nominal 1K input impedance to the modulation bus. Since the sub-carrier isolation resistances normally range from 10K to 200K each, the summing ratio is quite satisfactory. The amplifier is terminated in a cathode follower to provide isolation from the transmitter. Over-all amplifier gain has been made variable by means of adjustment of the feedback ratio over a limited range in order to provide versatile coverage of factors including differing numbers of sub-carriers feeding the bus, the particular pre-emphasis curve found desirable, and slight differences in transmitter deviation sensitivity. Normal over-all voltage gain normally lies in the region of 0.2 to 2.7.

II. Telemeter Transmitter

The Lockheed telemetering transmitter is rather unique in the airborne FM/FM field. Originally conceived as a design embodying a significantly improved sub-carrier-to-microphonic noise ratio over that of conventional methods, it utilizes some of the distinct advantages of both the self-excited modulator and a crystal-controlled reference frequency.

The usual self-excited transmitters are capable of linear, wide band deviation with low distortion but ordinarily are prone to microphonic sensitivity and suffer random frequency drift. Crystal-controlled transmitters have excellent frequency stability, of course, and the crystal stage is relatively free from microphonic disturbances, but the modulator is typically frequency sensitive, and is unable to produce wide-band deviation of the crystal stage without serious distortion. Coupled with this is the disadvantage inherent in the multiplication of spurious deviations caused by microphonism from the modulator up to the carrier frequency.

The Lockheed transmitter produces full-band deviation of ±125 KC in the first, or modulator, stage, which is operated in the low 1.2 to 1.5 MCS range and possesses a low microphonic figure. All subsequent operations in the transmitter serve to translate, or heterodyne the modulator output up to the telemetering 215-235 M.C. band, and then to provide sufficient power for radiation or further amplification purposes.

The modulator stage is comprised of an extremely stable R-C multivibrator which is frequency modulated by injection of the composite signal into the common grid return of both tubes. Output signal for frequency translation is capacitor-coupled from one plate. Modulator frequency excursions of 125 KC amount to 10.5% of center frequency for the lowest used, and are accomplished with a linearity of approximately 1% in a total bandwidth of 300 KC. Deviation sensitivity lies in the range of from 1.5 to 3.0 VRMS for ±125 KC, and is nominally 2.0 VRMS or 62.5 KC per volt, RMS, 44.5 KC per volt peak. Frequency sensitivity of the modulator, denoted by the input voltage needed to maintain constant deviation with various input frequencies, is essentially flat within the frequency response of the transmitter of 60 CPS to 80 KC between the 3 db points. Drift stability of the modulator is strengthened by the use of premium circuit components and a thorough aging process. External voltage control for the multivibrator is also furnished.

The first frequency translation is accomplished by mixing the modulator spectrum with a reference signal from a crystal-controlled oscillator operating in the 45 MCS region. Simple filters remove the unwanted modulation products passed by tuned mixer stage, and the modulator fundamental spectrum plus the crystal component is next passed to the 2nd mixer, which operates at the transmitter final frequency. In this stage, the nominal 47 MCS spectrum delivered by the 1st mixer is amplitude modulated by the 180 MCS-region signal derived from the 4th harmonic of the crystal-controlled frequency in the quadrupler stage.

With the 2nd mixer output at final transmitter frequency in two translations, a straight amplification process is then followed. The driver stage, utilizing a subminiature 5702 pentode delivers 1/2 - watt to the final amplifier, a single sub-miniature 5703 triode connected for grounded-grid operation. This final stage delivers 2 watts into 50 ohms in the range of 215-235 MCS.

Over-all transmitter characteristics of interest include:

1. Modulation total distortion of 1%, when considered separately from source and receiver distortion.

2. Maximum spurious deviation of 1.5 KC or 1.2% of the normal 125 KC deviation limit when subjected to 10g's peak vibration of from 20 CPS to 2000 CPS in any of three normal planes or to impact of 100 g's peak for a 3 millisecond duration.

3. An over-all frequency stability restricted to a maximum of ±0.02% of assigned operating frequency, but held to better than ±0.01% in practice with satisfactory airborne voltage regulator operation.

4. Weight: 1.7 lbs.
   Size : 3.6 x 6.1 x 1.6 inches
   Volume enclosed: 35.1 cubic inches
   Cooling: Conduction through mounting plate.

5. Input Requirements: nominal 2.0 VRMS across 15K OHMS
   Power Output: 2.0 watts into 50 ohms @225 MCS
   Input Power: (A+) 6.3 volts @ 1.5 amps or 9.45 watts
   (B+) 250 volts @ 90 M.A. or 22.50 watts.
   Total power: 22.50 + 9.45 ≏ 32 watts.

IV. Additional Power Amplification

In cases where flight range problems and the assurance of a high signal-to-noise ratio available to the receiver becomes important, straight class "C" power amplification of the transmitter output is accomplished by a tandem power amplifier used for either relatively high output power levels (50-70 watts) or for a moderate 10-watt level.

V. High-level power amplifier

The highly efficient Lockheed power amplifier is designed to secure a single stage power level of from 50 to 70 watts with an input of 2 watts, or a power gain of from 25 to 35 (a 15 db increase). Cylindrical in shape, it is 6 inches long and 3 inches in diameter, enclosing a volume of approximately 42.5 cubic inches, and weighs approximately 3 pounds. Indicated power per unit volume ratio is therefore approximately 1.2 watts per cubic inch in the 50-watt region alone. Higher power outputs can be obtained with an increase in transmitter driving power or with lighthouse tube plate voltages higher than the presently used 500 VDC. Cooling is provided with an external 1.5 inch centrifugal blower at 10,000 RPM.

VI. Moderate-Level Power Amplifier

The X-17 missile required only moderate power output and therefore utilized a nominal 10 watt output power amplifier derived from a self contained, conduction-cooled amplifier designed around the special 6030 double-tetrode power amplifier. Unlike the high-power cavity amplifier, this unit utilizes conventional grid and plate tank circuitry, operating into 50 ohms at both ends. A minimum of 1.4 watts driving power is needed to secure full rated output, but with the elimination of cathode self-biasing, 12-13 watts can be obtained with as little as 1.5 watts input power.

Power Requirements: A+ : 0.85 amps @ 6.3 volts (5.4 watts)
B+ : 110 MA. @ 250 volts (27.5 watts)

Size: 4.9 x 3.37 x 2 inches or 33.02 cubic inches.

Weight: 16 ounces or 1 pound

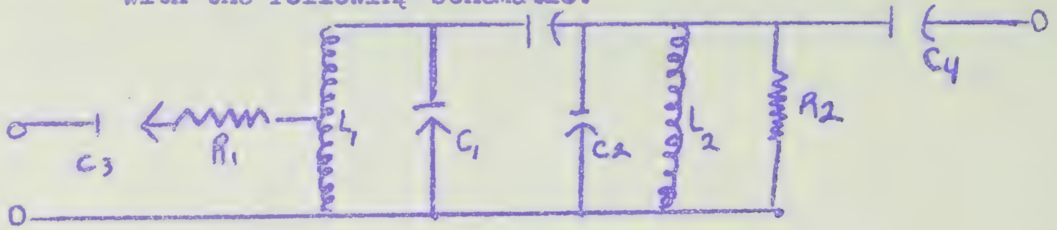
Minimum Vibration Specification: 10 g's peak from 20 CPS to 2000 CPS in 3 normal planes.

VII. R.F. Power Monitors

In order to gain data on possible changes in effective antenna loading caused by an ionization effect during the serious re-entry phase of flight, an R.F. monitor circuit was utilized to detect VSWR changes in the transmission line between the R.F. power amplifier and the vehicle antenna. A simple, low-volume assembly consisting of a potted coupling loop, a diode detector, and a resistance-capacitance filter, all totaling approximately one inch in length, was inserted into a machined coaxial tee fitting which became part of the transmission line itself. Maximum output voltage for telemetering monitoring approached 1.5 volts d.c. Linearity of the device was essentially linear over the range of from 5 to 12 watts and the sensitivity ratio approximated 0.2 VDC per watt reduction in power.

3.2.2 Detail Design and Construction:

3.2.2.1 Circuit Design: The circuit design shall be in accordance with the following schematic:



R_1 = 100 K ohms
$C_1 = C_2$
$L_1 = L_2$

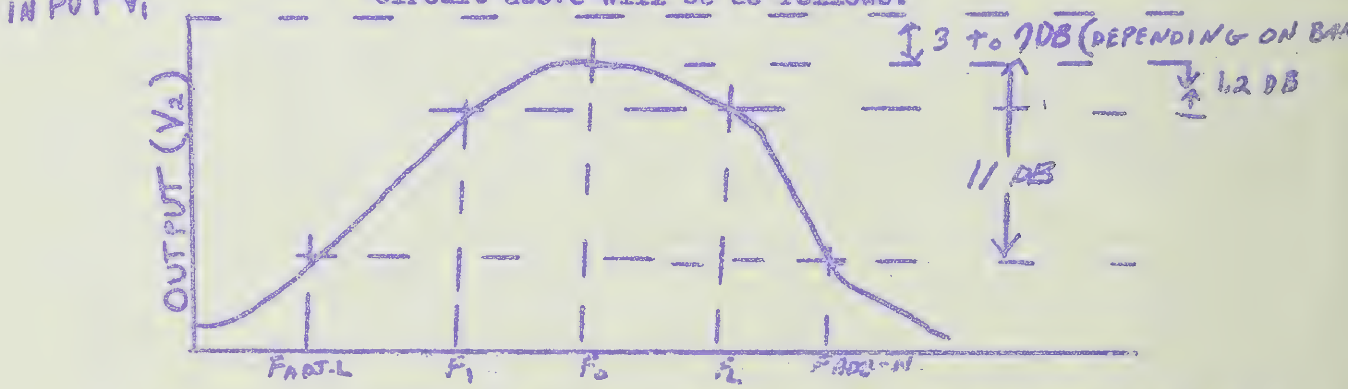
$Q_1 = Q_2$ = 7.2 ±10%
C_3, C_4 = DC Blocking

$k = k_c$
Where
$k = \frac{C_o}{C_1 + C_o}$

$k_c = \frac{1}{Q_1}$

NOTE: The capacitance "pi" made up of C_0, C_1, C_2 may be replaced by its equivalent "tee" if desired.

3.2.2.1.1 Response Curve: The typical response curve for the circuit above will be as follows:



NOTE: The allowable response limits are specified in 3.3

3.2.2.2 Construction: All units shall be embedded in a suitable thermosetting resin. A close fitting metal case shall be provided as specified on the applicable LMSD drawing.

3.2.2.2.1 Special Shielding: When the metal case specified does not completely enclose the filter, any remaining surfaces shall be provided with a temporary close fitting shield made of .010 thick cold rolled steel to complete the enclosure. This shield is necessary to insure performance characteristics in accordance with this specification.

3.3 Performance

3.3.1 Center Frequencies: Center frequencies shall be as specified on the applicable LMSD drawing.

3.3.2 Input Signal: The maximum input signal shall be 15 volts rms. with a 2000 ohm a-c source impedance. D.C. blocking must be provided at both input and output. This may be accomplished by the use of paper or ceramic capacitor of sufficient size to avoid alteration of the filter performance.

3.3.3. Output: To supply a load of 3 megohms shunted by 15 mmf of capacitance:

3.3.4 Attenuation: Attenuation at adjacent band edges shall be 10 db or greater as measured from peak response at the output terminal. Any harmonic response above the second harmonic of F_o and extending to 125KC shall have a minimum attenuation of 25 db below the attenuation at F_o.

3.3.5 Variation: Variation in response over the Frequency band (F_1 to F_2) shall not exceed 1.5 db as measured at the output lead. If the filter is tuned such that $k > k_c$ the resulting dip shall be no more than 0.5 db below the peak response. (See exception of 3.4.1)

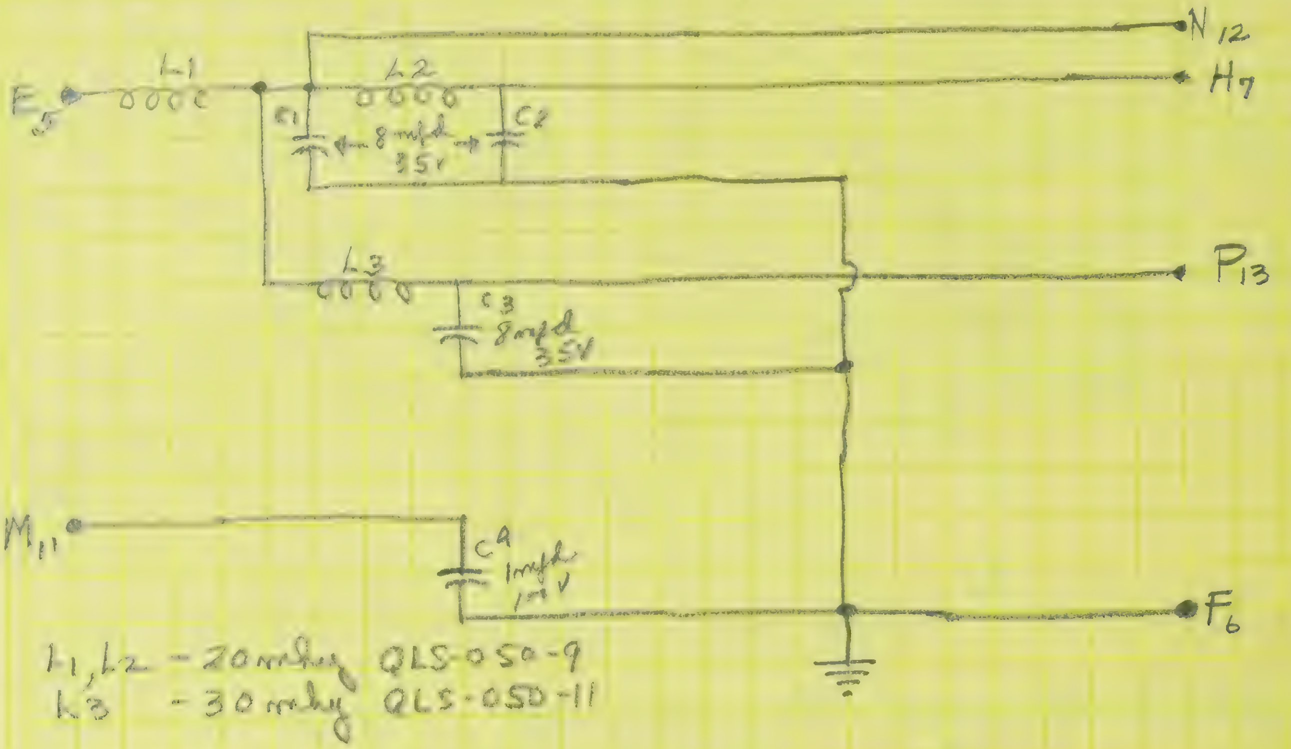
3.3.6 Maximum Loss: Maximum loss measured by the ratio of V_2/V_1, shall not exceed 3db on bands 10 thrugh 18, 5db on bands 6 through 9, or 7 db on bands 1 through 5. at peak response of the output. (See exception of 3.4.1).

3.3.7 Band Limits: Frequencies in cycles per second.

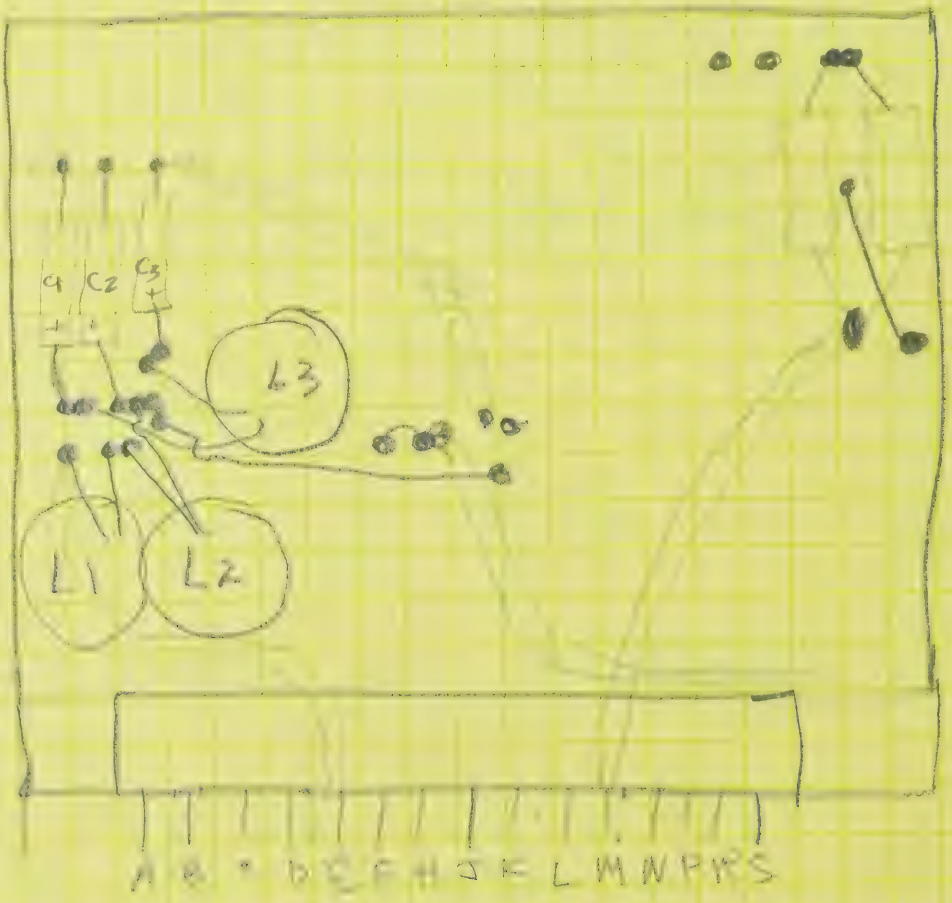
Band No.	Center Frequency	Pass	Band	Adjacent Edge of Next Lower	Adjacent Edge of Next Higher
	F_o	F_1	F_2	Band FABJ-L	Band FADJ-H
1	400	370	430		518
2	560	518	602	430	675
3	730	675	785	602	888
4	960	888	1,032	785	1,202
5	1,300	1,202	1,398	1,032	1,572
6	1,700	1,572	1,828	1,398	2,127
7	2,300	2,127	2,473	1,828	2,775
8	3,000	2,775	3,225	2,473	3,607
9	3,900	3,607	4,193	3,225	4,995
10	5,400	4,995	5,805	4,193	6,799
11	7,350	6,799	7,901	5,805	9,712
12	10,500	9,712	11,288	7,901	13,412
13	14,500	13,412	15,588	11,288	18,700**
14	22,000	20,350	23,650	15,588	27,750
15	30,000	27,750	32,250	23,650	37,000
16	40,000	37,000	43,000	32,250	48,560
17	52,500	48,560	56,440	43,000	64,750
18	70,000	64,750	75,250	56,440	-

** Upper adjacent band edge listed for band 13 is required for intermediate band not listed.

XA Filter

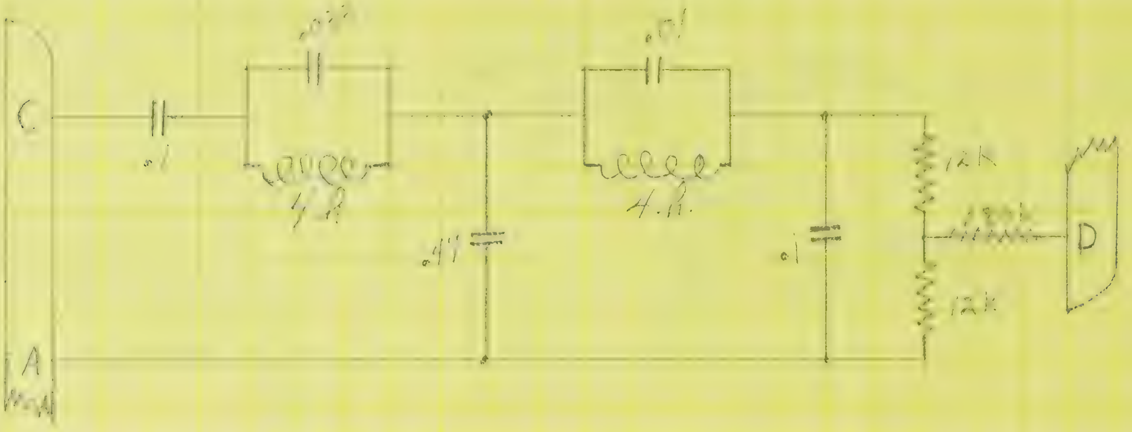


L1, L2 - 20 mhy QLS-050-9
L3 - 30 mhy QLS-050-11



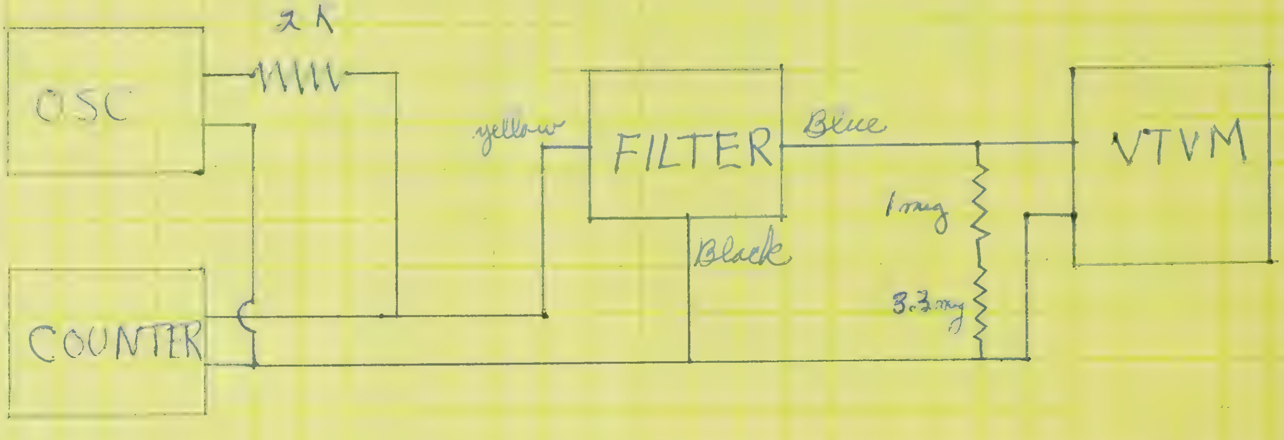
Low Pass Filter

1034542
#Spec #10530



1. Peak response at 230 ± 50 cycles
2. Insertion loss at peak response less than 20d
3. Cutoff - output shall be down a minimu
of 40 db from 600 cps to 3K cps
4. Attenuation - output at 150 cps shall not
be down more than 10 db.

FILTER TEST



1. Set VTVM to 10 volt range.

2. Tune oscillator to peak (around F_0) and set input for indicated scale reading of "0" DB.

3. Short blue and yellow leads. Read increase in DB. (3DB or less for ch. 10-18, 5 DB for ch. 6-9, and 7DB for ch. 1-5).

4. Vary oscillator between F_1 and F_2. Output variation should not be more than 1.5 DB

5. If dip is noticed near F_0; the amplitude of the dip should be within .5 DB of peaks

6. Tune oscillator to adjacent channel edges. Attenuation should be at least 10 DB. below max.

7. Tune oscillator to twice F_0. Attenuation should be at least 25 DB.

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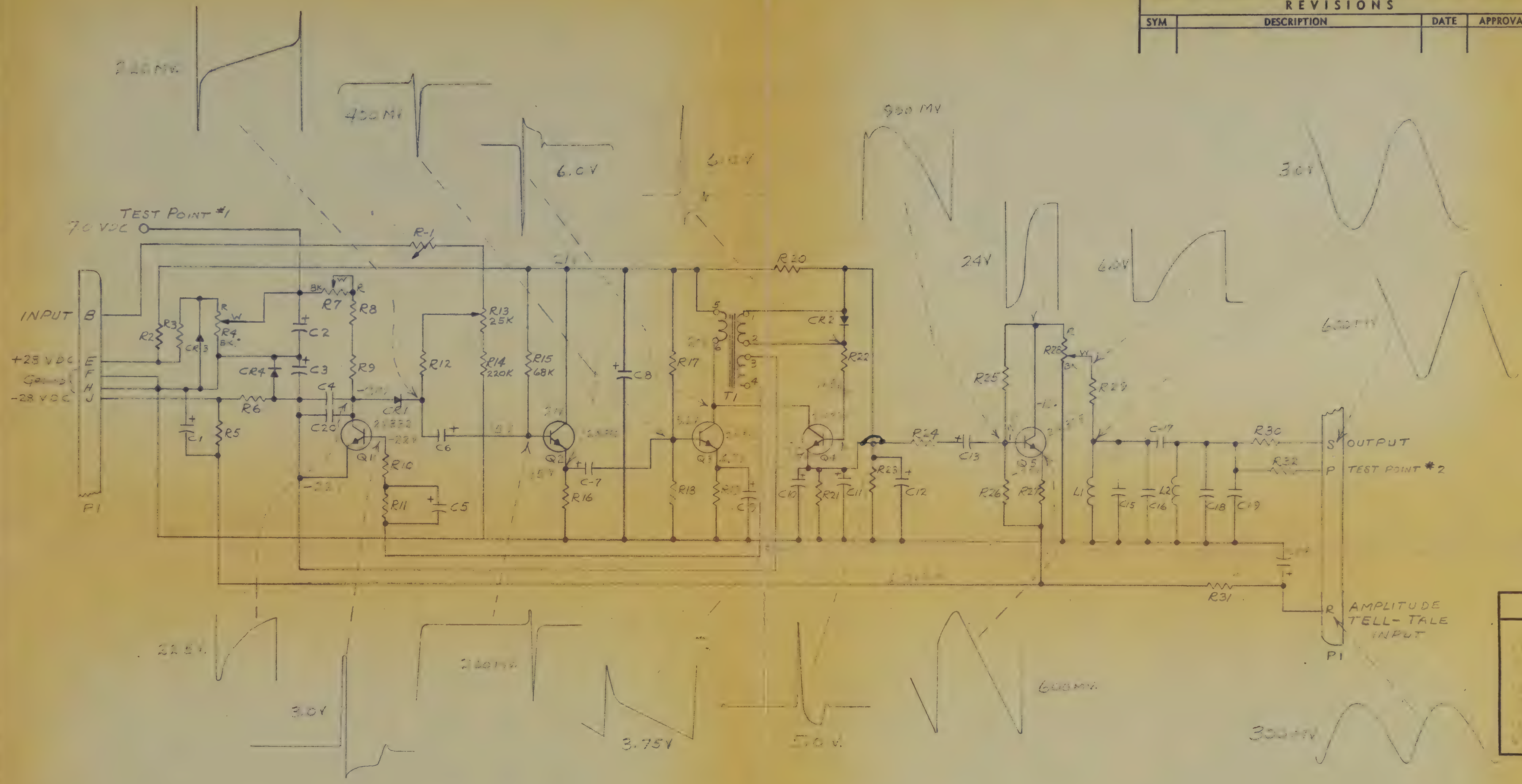
REVISIONS

SYM	DESCRIPTION	DATE	APPROVAL



9. VOLTAGES & WAVEFORMS TAKEN FROM A CHAN. 11 OSC.
8. ALL WAVEFORM VOLTAGES ARE PEAK TO PEAK VOLTAGES
7. R24 – 2KΩ [illegible]
6. R28 IS 10K POT TO SET OUTPUT VOLTAGE.
5. R24 - NEVER BELOW 2KΩ, USED FOR OUTPUT VOLTAGE.
4. R13 IS 25K POT. FOR INPUT SENSITIVITY.
3. [illegible]
2. R4 IS 20K POT. TO SET 7.0 VDC TEST POINT VOLTAGE.
1. ALL VOLTAGES TAKEN WITH A SIMPSON MODEL 260 METER EXCEPT 7.0 VDC TEST POINT.

	PART NO.	ZONE	DESCRIPTION	MATL	MATL SPEC	NEXT ASSY	USED ON	NEXT ASSY	FINAL ASSY
REQD PER ASSY	LIST OF MATERIAL					APPLICATION		QTY REQD	

UNLESS OTHERWISE SPECIFIED DIMENSIONS ARE IN INCHES
TOLERANCES
DECIMALS .X ± .XX ± .03 .XXX ± .010 ANGLES ±
HEAT TREAT
FINISH

ORIGINAL DATE OF DRAWING [illegible]
DRAWN [illegible]
CHECKED [illegible]
APPROVAL
CUSTOMER
DRAWING SIZE C

TITLE
[illegible] WAVEFORMS AND VOLTAGES AT [illegible]

SCALE
WT ACT. LBS WT CALC LBS

LOCKHEED AIRCRAFT CORPORATION
MISSILE SYSTEMS DIVISION
VAN NUYS, CALIF.

SHEET OF

MISSILE SYSTEMS DIVISION

INTERDEPARTMENTAL COMMUNICATION

TO Distribution DATE March 18, 1958

FROM L. E. Hamilton DEPT. 51-46 PLANT B-9 EXT. 1060

SUBJECT: 12-204 STRAIN GAUGE MEASUREMENTS

I. PURPOSE OF REPORT

This report is for the following purposes:

1. To describe a technique that is to be used for the measurement of strains as low as 60 micro-in/in, in the 12-204 Polaris and the errors that can be expected.
2. To describe a system design which has been developed and tested and is capable of giving excellent results (system was tested on RUB Channel 13).
3. To provide specific information about the proper techniques to be used in setting up the electronic system.

II. SCOPE OF REPORT

It is intended that this report shall serve:

1. As a basis for the establishment of new procedures for set-up and check out.
2. As a reference document for the proper means of providing last minute range changes.
3. As a reference document in cases of system mal-functions.
4. As a means of assuring that calibration data is properly recorded.

III. TECHNIQUES FOR LOW RANGE STRAIN GAUGE MEASUREMENTS

Figure 1 shows the normal per channel telemeter system using a high gain dual loop oscillator and a four active arm strain gauge bridge. The maximum sensitivity for this system is about 1000 μ in/in.

The problem is, therefore, to increase the maximum sensitivity of a telemeter system employing a dual loop subcarrier oscillator to give full band width, with 60 μ in/in strain in two active arms of a 4 arm bridge.

Since it was not possible to reliably increase the side loop gain of the high gain subcarrier oscillator a small external transistorized amplifier was developed and placed in the side loop as shown in Figure 2.

In addition, the high gain dual loop oscillator was partially re-designed using, improved transformers, better wiring techniques to reduce pickup, plate voltage decoupling; thereby, eliminating the necessity of two 150 volt supplies, an improved phase shifting network in the side loop, and provision for both balanced output and input.

60 μ in/in full scale sensitivity represents a resistance change of .012% for a two active arm strain gauge bridge having a gauge factor of 2.

It is obvious that for a full scale change of .012% the initial balance of the bridge must be very good. In addition to an extremely good resistance balance, an extremely good reactance balance must be made to insure that no quadrature components of the input voltage are fed into the side loop of the dual loop oscillator.

Because of the extreme sensitivities and resulting balancing problems, the two inactive arms of the bridge are strain gauges of the same type and are bonded to the same type of metal as the two active arms. This approach helps minimize the temperature and capacity problems since these effects should be approximately the same in all 4 arms and therefore give no net unbalance.

Since the two active arms are separated some distance from the two inactive arms and must be connected by the means of a cable, temperature effects in the cable can give large apparent strains. It should therefore be apparent, that there must be equal lengths of copper wire in all four arms if these effects are to be eliminated. Figure 3 shows the wiring diagram that MUST be followed. Cables 1, 2, 3, and 4 must be exactly the same length if good temperature compensation is to be obtained. It can readily seen that if cables 1, 2, 3, and 4 are equal there are equal lengths of wire in all 4 arms.

If the installation will permit, it is highly desirable to have cables 1 & 3, 2 & 4, 6 & 8 twisted pair shielded. (NO OTHER COMBINATIONS ARE PERMISSABLE). This will help eliminate undesireable effects that can be caused by variable spacing of the cables.

The cabling shown in Figure 3 is for an installation where the two active arms are on opposite sides of the skin. For this type of installation, the outside cables will be bonded to the skin. For good temperature compensation, the cable on the inside should also be bonded to the skin and be approximately opposite to the outside cable.

Best temperature compensation has been obtained with cables 1, 2, 3, and 4 inside a copper braid jacket. Using a composite cable of this type between the active and inactive arms, it has been possible to imbed 10 feet of cable in dry ice with negligible effects, (less than 1%).

Figure 3 also shows the location of 4 calibration impedances Z_1, Z_2, Z_3, Z_4, that are necessary to completely balance the bridge. In general, these impedances are resistors in parallel with small capacitors.

It has been found that only resistors of the deposited carbon type are suitable for use any place in the bridge or calibration circuits. It has been found that usually only two resistors and one capacitor are necessary to balance the bridge.

STRAIN GAUGE AMPLIFIER:

The strain gauge amplifier installation has been so designed that the amplifier may be removed from the circuit and the bridge output connected directly to the dual loop oscillator without the necessity of re-phasing, should the sensitivity requirements of the system be above 1000 μ in/in.

To meet this requirement, it was necessary to add the resistor R_{c} in series with the output circuit as shown in Figure 3.

Since the input impedance and gain must be constant if the amplifiers are to be interchangeable, lab sets have been provided internally to accomplish this.

GAIN CHANGES:

All oscillators and amplifiers will be set in the laboratory for the same system gain of 60 μ in/in. Should the requirements change to a higher value the sensitivity can be easily reduced in the field by shunting the bridge with the resistor R_{c3} as shown in Figure 3.

As can be seen this reduces the sensitivity of only the bridge and does not affect the in-flight calibrate circuit, R_{c1}, R_{c2}, R_{1}, R_{3}.

IV. SYSTEM PERFORMANCE

In the laboratory, it has been possible to consistently get system performance, linearity, repeatability of less than 2% and in most cases less than 1%. The temperature stability of the electronic components has been excellent.

A test sample, consisting of a known area steel bar with the two active arms bonded to it, has been subjected to known strains. The resulting curve of strain Vs. frequency shift has been always linear and repeatable to an accuracy of better than 1%. Deflections as small as .22 μ in/in were detectable. It has also been possible to make the system work with a full scale sensitivity of 33 μ in/in.

In addition, the test sample was immersed in salt water at room temperature and bubbles blown through the water past the sample. The output of the oscillator was fed into a subcarrier discriminator and its output recorded. The effect of the bubbles (which could show up as excessive noise) was negligible.

The strain gauges on the test sample were bonded to the plate, and had an electrostatic shield over them. The entire assembly was waterproofed with a pliable waterproofing material. Since salt water is conducting, this type of installation MUST be used for all active strain gauges. The waterproofing must be of high enough quality so that at least 500 megohm insulation resistance is maintained at all times and does not deteriorate after prolonged exposure to adverse atmospheres. Test units showed a resistance of 1000 megohms after salt water immersion.

It is assumed that the strain gauges to be used on the vehicle will be carefully selected and installed in such a manner that the change in resistance due to thermal effects is negligible.

V. SYSTEM LAYOUT

Figure 4 is a schematic of the overall system as to be installed in the test vehicle showing the important areas of access.

The interconnect area will contain the components shown in dashed area of Figure 3. THIS AREA MUST BE EASILY ACCESSIBLE DURING SET UP. Not shown in the schematic but to be included in the package are adequate terminal boards for mounting the balancing components Z_1, Z_2, Z_3, Z_4.

Since all the amplifiers are closely spaced, care should be taken to have adjacent amplifiers separated in frequency.

For elimination of undesireable capacity effects, it is highly desireable to have installations with long cable runs on the lower channels.

OPERATIONAL:

In view of the stringent requirements imposed by the high sensitivity of the measurements, all possible means must be utilized to eliminate the need for extensive set up in the final installation. To achieve this end, the design was tailored about the following mode of operation:

1. Standardization of the oscillator and amplifier performance at set up.
2. Adjustment of only the bridge balance and full scale range in the final installation.
3. Calibration resistors of one known value with a final check of system performance by shunting a known value resistor across the inactive arms of the bridge.

VI. SET UP

SUBCARRIER OSCILLATOR (DRAWING NO. 1034095)

The balance and phasing of the output transformer should be checked by disconnecting the bridge and connecting resistors R_1 and R_2 in its place (R_1 and R_2 are a pair of 175 ohm deposited carbon resistors accurately matched) a VTVM is connected between point A and ground and the voltage read. It should read less than .01 volts. Reversing the resistors should make no significant difference in the reading.

The subcarrier oscillator should be adjusted in a circuit as shown in Figure 5. (Omit R_1 and R_2 for the following tests). The 4 gauges should be properly balanced using the plan outlined under check-out and should be bonded to a thick plate. The gauges should accurately simulate the flight items. Full channel deviation with minimum amplitude shift should be obtained with 1000 μ in strain as simulated by the calibration resistors R_{c1}, R_{c2}. The strain simulated by the calibrate resistors is given by the formula $S = \frac{R_g}{G.F.\ R_c}$ μ in.,where

R_g is the gauge resistance in ohms, G. F. is the gauge factor of the gauges and R_c is the calibrate resistance in megohms. R_{14} and R_3 are 220 and 10K ohms, respectively. If R15 is not approximately 220 Ω for ± 1000 μ in/in full scale sensitivity, the oscillator should be thoroughly checked for defects. The amplitude shift should be held to standard tolerances and is adjusted by the means of C15. Due to the low levels involved, shielded cable and short wires should be used in the fabrication of the test circuit.

STRAIN GAUGE AMPLIFIER (DRAWING NO. 1033987)

Four things must be known if the strain gauge amplifier is to perform in the system. These are:

1. The input impedance MUST be constant. This factor is more important than gain or phase shift since it must be constant and known if the operational plan of shunting the bridge to reduce sensitivity is to work and only one value of calibrate resistor is to be used.

2. The gain should be preset to a known value.

3. Phase shift should be checked.

4. Common mode gain should be essentially zero.

At this time, it is planned to channel set the amplifiers. It is hoped that further tests will indicate that this is unnecessary.

In view of the extremely low levels that the amplifier is to operate at (approximately 120 μ volts in, 3 MV out) normal methods of measurement are unsuitable because of the noise problem.

To eliminate the undesireable noise effects, that would be present using normal measurement methods, the circuit in Figure 6 should be used. The 4 strain gauges MUST be properly balanced using the plan outlined under check out and MUST be bonded to a thick plate. The calibration resistors are of such value to simulate 60 μ in/in strain and are calculated as outlined above under "Subcarrier Oscillator".

1. BIAS With the dual loop oscillator out of the circuit and a VTVM on the emitter of Q2, adjust R_2 for a reading of 2.45 Volts D.C. with 6.8 Volts, B+. (IMPOSSIBLE. AMP. MUST HAVE A LOAD)

2. With a dual loop oscillator of the proper channel, set up as outlined above, and K_1 and K_2 closed, adjust R_6 for full channel deviation when the cal + and cal - circuit is operated. R_6 should be approximately 450/550 ohms; if not, check the unit for defects.

3. Check the common mode gain by closing K_3 and K_4. The deviation obtained by closing K_3 and K_4 should not exceed 15 cps.

4. With K_1 and K_2 closed, operate cal + and cal -, recording the deviation. With K_1 and K_2 open again, operate cal + and cal -, recording the deviation. Compute the ratio of deviations and find the input impedance using the chart given in

Figure 7. Adjust R_8 and repeat the test until the input impedance is 600 ± 20 ohms.

5. Check the amplitude shift for full channel deviation. It should be less than 5%.

NOTE: GAIN AND INPUT IMPEDANCE ARE RELATED. ALWAYS RE-CHECK THEM AFTER ANY ADJUSTMENTS ARE MADE. ALL FINAL CHECKS MUST BE MADE WITH THE COVER ON.

VII. CHECK OUT

With the completion of the above set up procedure, bench check out of the system is unnecessary except to insure proper wiring.

During final installation, the following must be accomplished:

1. Determine operability of oscillators, amplifiers, and calibrate circuits using cal 0, cal +, and cal -. UNTIL THE BRIDGE IS BALANCED THE SYSTEM PROBABLY WILL NOT WORK EXCEPT ON THESE POSITIONS.
2. Balance the bridge using the procedures outlined below.
3. Operate cal + and cal - circuits and record deviations and amplitude shift.
4. Shunt the standard carbon film resistor R_s, across the inactive arms of the bridge and note deviation; this should agree with the deviation obtained using calibrate circuits. Check the calibrate resistors, if necessary.
5. Using the graph in Figure 5, select the proper shunt resistor to reduce the sensitivity to the desired value. After installation shunt inactive arms with the proper carbon film resistor and note deviation.

BRIDGE SET UP

The most difficult problem in the entire system is the balancing of the bridge. The following procedures have yielded good results and rapid set up. The test equipment necessary to balance the bridge is shown in Figure 9.

1. Connect the apparatus as shown in the diagram with the properly terminated channel filter connected as shown.
2. Ground the input of the oscillator as shown.
3. Connect R_7 and R_8 to either terminal 3 or 4. Connect the probe to point 4 and adjust the phasing network for 0 phase shift on the oscilloscope, then disconnect R_7 and R_8.
4. Connect the probe to either pin 1 or 8 of the amplifier input and select a carbon film resistor that gives minimum indication on the CRO when connected between some pair of terminals 1, 2, 3, 4. In general, the minimum indication will not be zero but a circle.

5. Connect the probe to the other input terminal and repeat step 4.

6. Connect the probe to the output (terminal E and readjust the resistors for minimum indication.

7. Select a capacitor that gives 0 phase shift when connected between some pair of terminals 1, 2, 3, 4.

8. Readjust resistors and capacitors until output is close to zero as indicated on CRO.

9. Remove the oscillator input ground and observe amplitude and frequency shift.

10. Make final adjustment of resistors and capacitors to bring frequency and amplitude shift with tolerances.

In general, it should be possible to balance the bridge with 2 sets of 3 resistors in series and 1 capacitor. The connection of the resistors and capacitor as shown is for illustration only and the actual connection can be between any combination of 2 terminals.

The resistance of either of the sets of resistors should be greater than 25K and the capacitor should not be greater than a few micro-microfarads.

For first approximations, a 100K Bourns trim pot (and a series resistor, if necessary) has proved invaluable. The pot later being measured and replaced with fixed resistors. A small trimmer capacitor was also proved invaluable.

NOTE: DURING FINAL INSTALLATION OF RESISTORS HEAT SINK MUST BE USED. ADEQUATE TIME MUST BE GIVEN FOR THE RESISTORS TO COOL BEFORE READINGS ARE TAKEN.

AMPLIFIER REMOVAL

For full band strains of 1000 μ in. or greater, the strain gauge amplifier may be removed from the system. When the amplifier is removed, the output resistor Ro must be shorted. The amplifier output ground must be removed so the oscillator will have a balanced input. Amplifier input and output may now be connected.

Strains can still be simulated by shunting standard resistors across the bridge. However, the calibration resistor is no longer constant and must be selected using a known simulated strain.

A technical bulletin containing the theoretical discussion of the problem and equations involved is now being written.

L. E. Hamilton
L. E. Hamilton
Research Engineer
Electronics Group 11

APPROVED: JAC
J. A. Cox, Group Leader

LEH:tdl

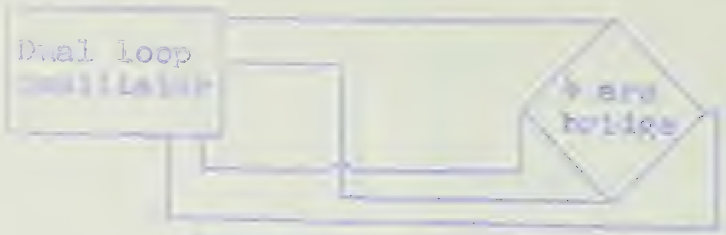


Figure 1

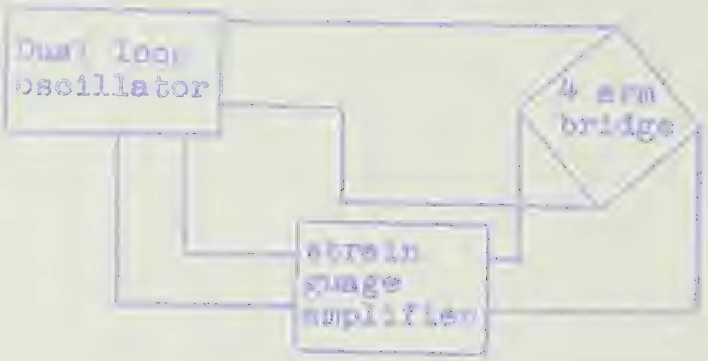


Figure 2

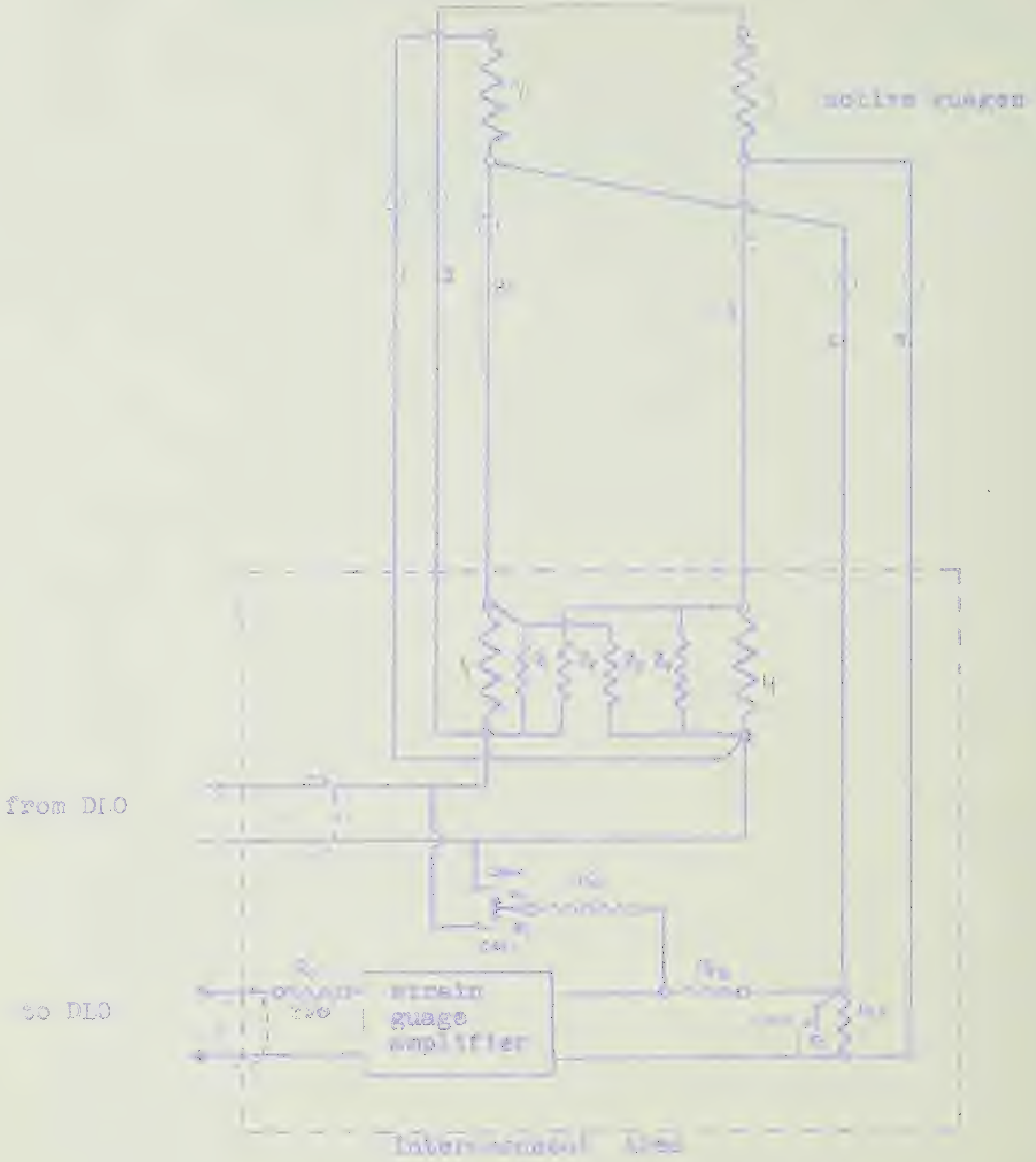

active guages
from DLO
to DLO
strain
guage
amplifier
Inter-connect box

Figure 3

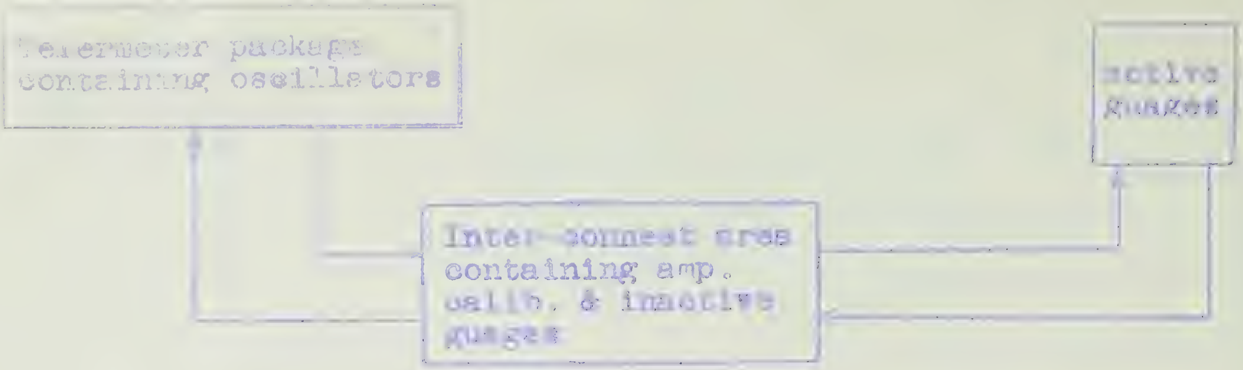

Telemeter package containing oscillators
Inter-connect area containing amp. calib. & inactive guages
active guages

Figure 4

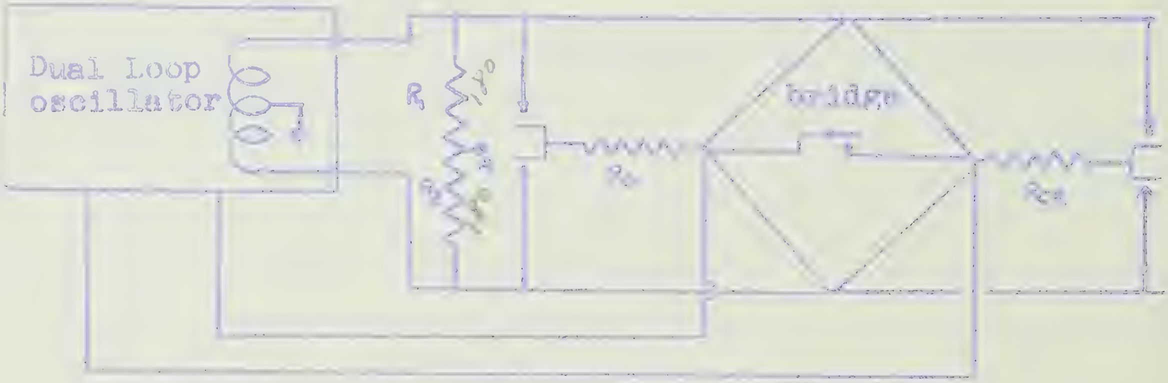


Figure 5

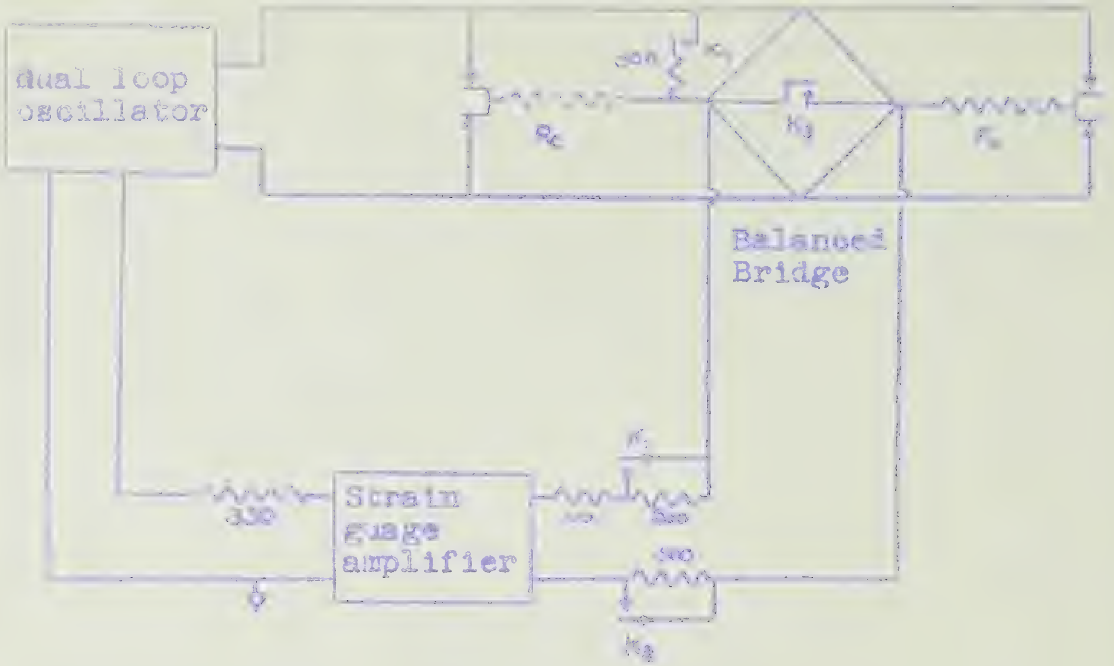

dual loop
oscillator
R_c
Balanced
Bridge
Strain
guage
amplifier
330

Figure 6

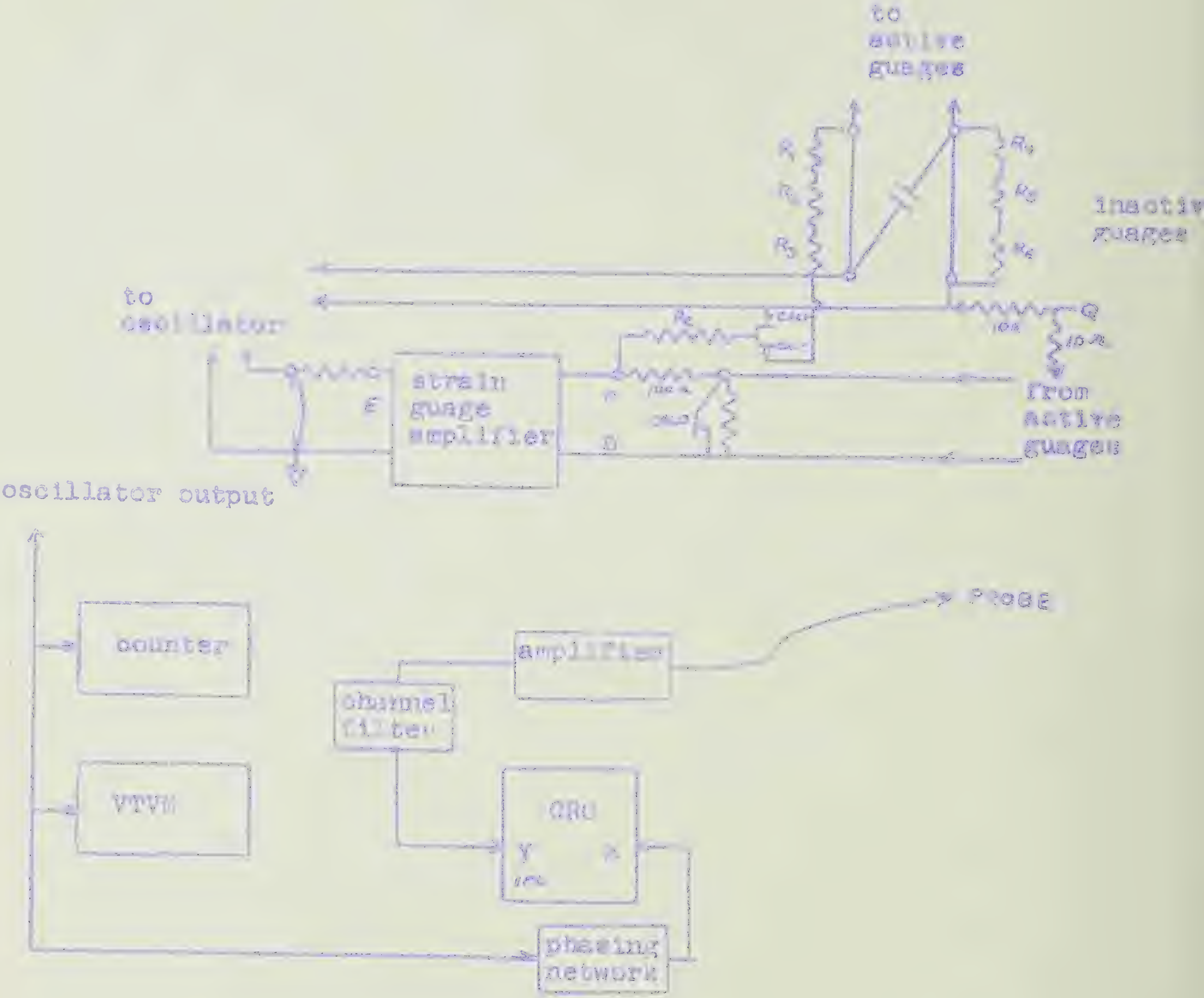

to
active
guages
R_1
R_2
R_3
R_4
R_5
R_6
inactiv
guages
to
oscillator
R_c
10Ω
10 Ω
Q
strain
guage
amplifier
E
D
from
active
guages
oscillator output
counter
VTVM
amplifier
PROBE
channel
filter
CRO
Y
X
phasing
network

Figure 9

Distribution:

E. C. Donahue
H. R. Francis
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W. Z. Meyer
W. S. Milenki
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J. F. Meerie
E. C. Veragge
File

	Name	Date	LOCKHEED AIRCRAFT CORPORATION MISSILE SYSTEMS DIVISION	1 Page		
Prepared	P E Hamlt	4/3/58	Title STRAIN GAUGE MEASUREMENTS FOR 12-204	Model		
Checked	P E Hamlt	4/3/58				
Approved		4/3/58		Report No 3457		

INSTRUMENTATION APPLICATION DEPARTMENT
ELECTRONICS GROUP 11

TECHNICAL BULLETIN

PURPOSE OF REPORT:

1. To discuss some of the problems encountered in the 12-204 Polaris strain gauge measurement in more detail than was possible in the IDC covering this subject.

2. To present useful mathematical formulas for calibration of such a system.

TECHNIQUES FOR LOW RANGE STRAIN GAUGE MEASUREMENTS:

An IDC titled 12-204 Strain Gauge Measurements by this author was issued on March 18, 1958 and should be consulted for general system information. Things that are obvious in the IDC will not be repeated here.

Figure 1 shows the normal per channel telemeter system using a high gain dual loop oscillator and a four arm strain gauge bridge. Figure 2 shows the equivalent circuit of such a system (See appendix for derivation of bridge circuit). Since the input transformer of the dual loop oscillator is terminated in the phasing network, the input impedance is NOT a pure resistance, but an impedance Z_{in}.

To increase the sensitivity beyond the maximum reliable sensitivity of ±1000 µ In/in possible with the dual loop oscillator alone, a small transistorized amplifier was added as shown in Figure 3. The equivalent circuit of such a system is shown in Figure 4. It has been possible to get full band sensitivities of ±33µ In/in with this system.

A desireable feature of such a system is the ability to go to lower sensitivities by removing the amplifier and jumpering its terminals without the necessity of any other adjustments (such as re-phasing the dual loop oscillator).

Since the dual loop oscillator is sensitive to any unbalanced phase shift in the feedback loop, it should be obvious from the two equivalent circuits that R_2 must be identical to R_2 if there is to be no change in when going from the first to the second (if Z_{in} is constant).

To meet the requirement, it was necessary to add a resistor in series with the output of the strain gauge amplifier since its intrinsic output impedance was lower than the output impedance of the bridge.

CALIBRATION CIRCUIT:

The complete circuit for the bridge and amplifier including the inflight calibration circuits is shown in Figure 5. Using the equivalent circuit for the bridge previously derived, yields the circuit shown in Figure 6 for normal operation.

It has been found that the input impedance of the strain gauge amplifier is mainly resistive therefore R_L can be substituted for Z_L.

From this circuit, it is obvious that $\frac{E_o}{E_{in}} = \frac{GF}{2} S \frac{Z_L}{(R_o + R_{o2} + R_L)}$ (1)

Let E_o' equal the voltage feed to the amplifier during calibration, as derived in the appendix:

$$E_o' = \frac{1}{2} \frac{R_p}{R_{c1}} \tag{2}$$

Equating (1) and (2) and solving for S

$$S = \frac{R_p (R_o + R_{c2} + R_L)}{G.F.\ R_{c1} Z_L} \tag{3}$$

Alternate Calibration Circuit

Since strain causes the bridge to change resistance, tension can be simulated by shunting the active arms with a resistance and compression simulated by shunting the active arms with a resistance.

Since:

$$\frac{\Delta R}{R_o} = G.F.S. \tag{4}$$

and the parallel combination of two resistors R_o and R equals $\frac{R_o R}{R_o + R}$ (5)

Let $\Delta R'$ = change in resistance due to shunting R_o with R

$$\Delta R' = \left[R_o - \frac{R_o R}{R_o + R} \right] = \frac{R_o^2}{R_o + R}$$

$$\Delta R' = \frac{R_o^2}{R} \text{ if } R \gg R_o \tag{6}$$

substituting (6) in (4) yields:

$$\text{apparent strain } (S) = \frac{R_o}{G.F.R} \tag{7}$$

$$\text{and } \frac{\Delta S}{\Delta R} = \frac{-R}{G.F.R^2} \tag{8}$$

SHUNTING OF BRIDGE TO REDUCE SENSITIVITY:

Last minute sensitivity changes can be easily made by shunting the bridge with a resistor R_{c3} as shown in Figure 5, with this resistor in, the equivalent circuit of Figure 6 changes to the one shown in Figure 7. Solving this circuit for E_o^* yields:

$$\frac{E_o^*}{E_{in}} = \frac{G.F.S}{2} \frac{R_{c3} R_L}{R_o (R_{c2} + R_{c2} + R_L) + R_{c3}(R_{c2} + R_L)} \tag{9}$$

However for full band deviation $\frac{E_o}{E_{in}} = \frac{E_o*}{E_{in}}$

Therefore equating (1) and (9) and solving for R_{c3} yields:

$$R_{c3} = \frac{R_o(R_{c2}' + R_L)}{\left[\frac{S*}{S} - 1\right](R_o + R_{c2} + R_L)} \qquad (10)$$

where $\frac{S*}{S}$ is the ratio of the full bandwidth strain with R_{c3} in the circuit to the full bandwidth strain with R_{c3} out of the circuit.

INPUT IMPEDANCE MEASUREMENT:

It can be readily seen from the foregoing equations that: the input impedance of the strain gauge amplifier MUST BE KNOWN.

The best method of measuring the input impedance of the amplifier is with it actually in the circuit (dual loop oscillator connected to a frequency indicating device). To do this, a known stimulus is applied and the deviation again noted. Since deviation is proportional to input voltage and the input voltage will be less with the series resistor in the circuit the input impedance can be computed.

Figure 7 is the equivalent circuit for measurement of input impedance.

With the switch closed:

$$\frac{E_o}{E_{in}} = \frac{R_L}{R_o + R_L} \qquad (11)$$

With the switch opened:

$$\frac{E_o'}{E_{in}} = \frac{R_L}{R_o + R_1 + R_L} \qquad (12)$$

Dividing (11) by (12) and solving for R_L

$$R_L = \frac{R_1}{\left[\frac{E_o}{E_o'} - 1\right]} - R_o \qquad (13)$$

Since the ratio of deviations of the dual loop oscillator is proportional to $\frac{E_o}{E_o'}$, $\frac{D}{D'}$, can be substituted in the formula.

APPENDIX

The matrix for a bridge as shown in Figure A is:

$$\begin{bmatrix} \dfrac{Z_n + Z_m}{Z_n - Z_m} & \dfrac{2Z_m Z_n}{Z_m - Z_m} \\[2ex] \dfrac{2}{Z_n - Z_m} & \dfrac{Z_n + Z_m}{Z_n - Z_m} \end{bmatrix}$$

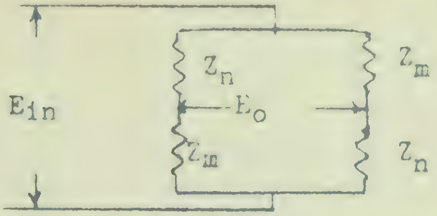


Figure A

If Z_n are strain gauges that change resistance due to strain, and this change is given by $\Delta R = G.F.\, R_o S$ where R_o is the intrinsic resistance of the gauge, G.F. the gauge factor and S the strain, and Z_m are inactive balancing gauges of the same type as the active; it can be easily shown, that the equivalent circuit of such a bridge is the circuit shown in Figure B.

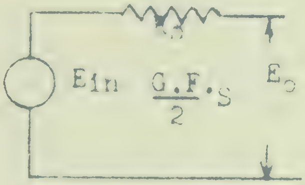


Figure B

Proof of Calibration Equation

Figure C is the circuit for either cal + or cal - (Polarity reversed for cal -)

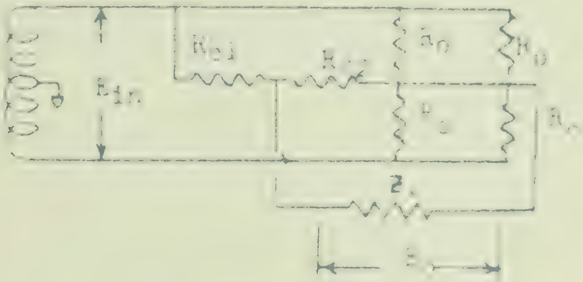


Figure C

Solution of the two loop equations yields $\dfrac{E_o}{E_{in}} = \dfrac{1}{2}\ \dfrac{R_p}{R_{c1} + R_{c2} + R_o/4}$

where $R_p = \dfrac{R_{c2}\ Z_L}{R_{c2} + Z_L}$

if $R_{c1} \gg R_o/4 + R_{c2}$

then $\frac{E_o}{E_{in}} = \frac{1}{2} \quad \frac{R_p}{R_{cl}}$

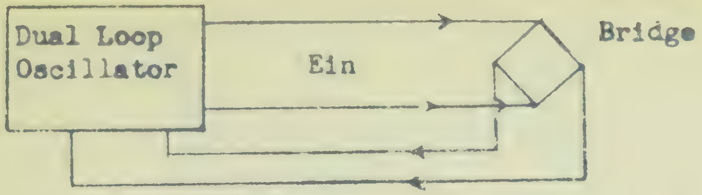
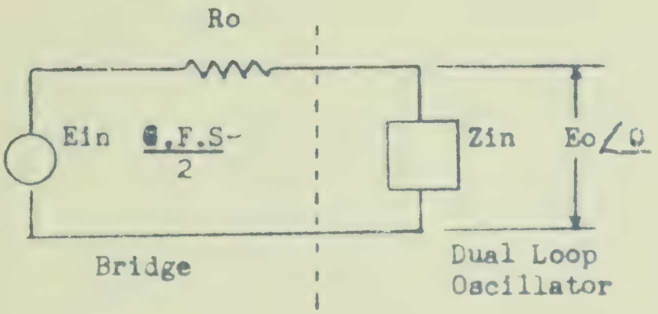


Fig. 1



Deviation = f [Eo, θ]

Fig. 2

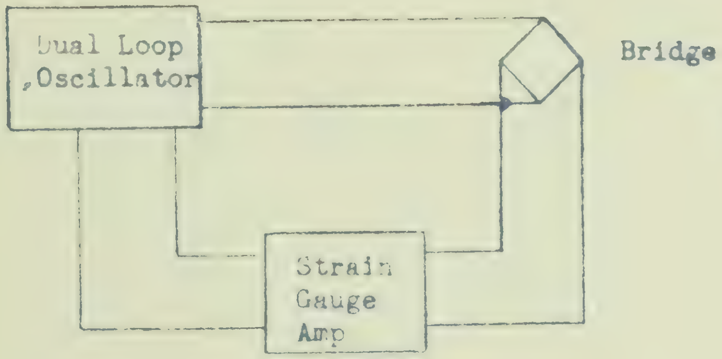
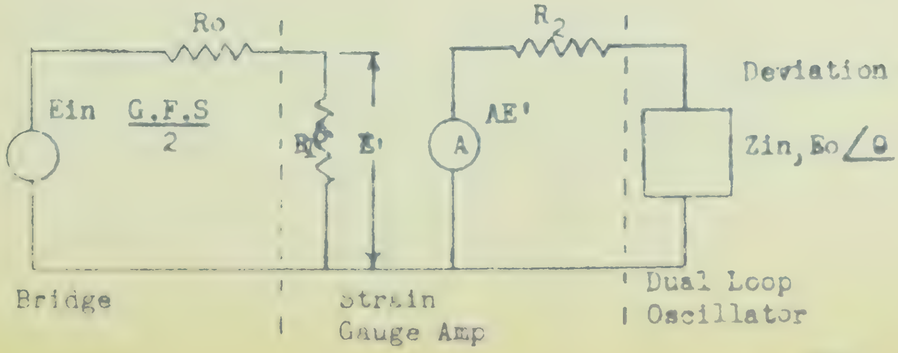
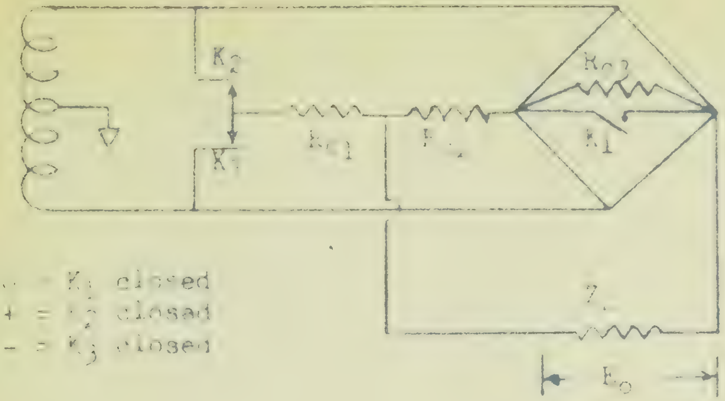


Fig. 3



Deviation = f [Eo, θ]

Fig. 4



Cal 0 = K_1 closed
Cal + = K_2 closed
Cal − = K_3 closed

Fig. 5

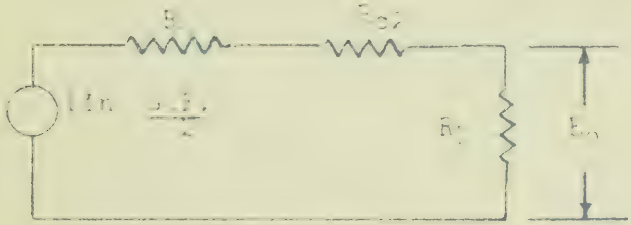


Fig. 6

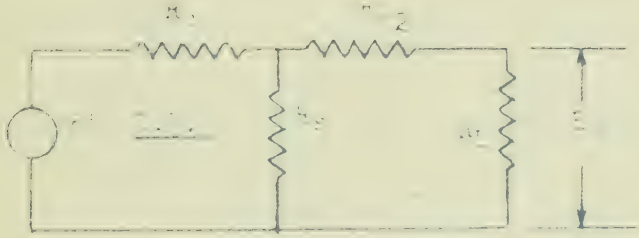


Fig. 7

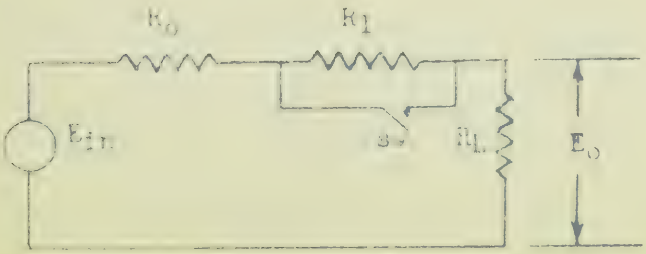
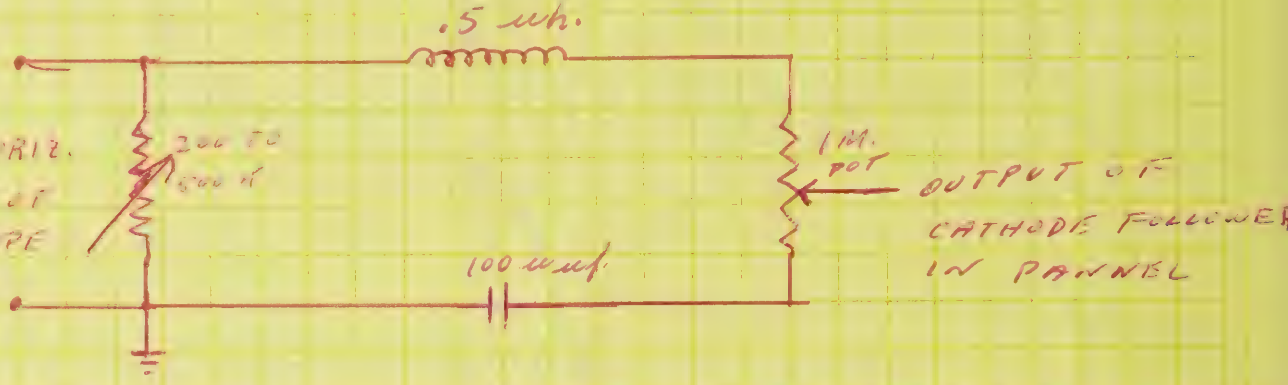
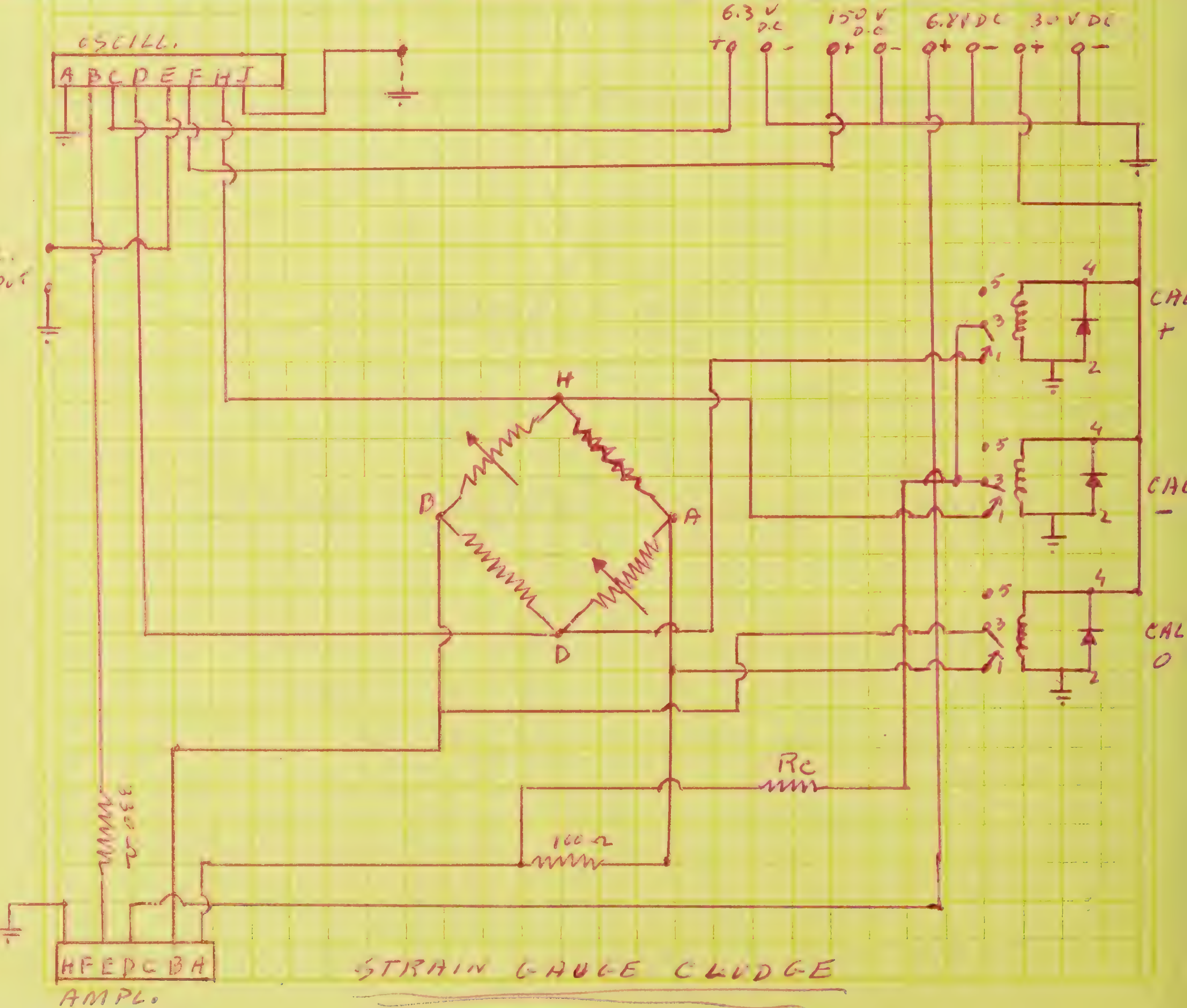


Fig. 8

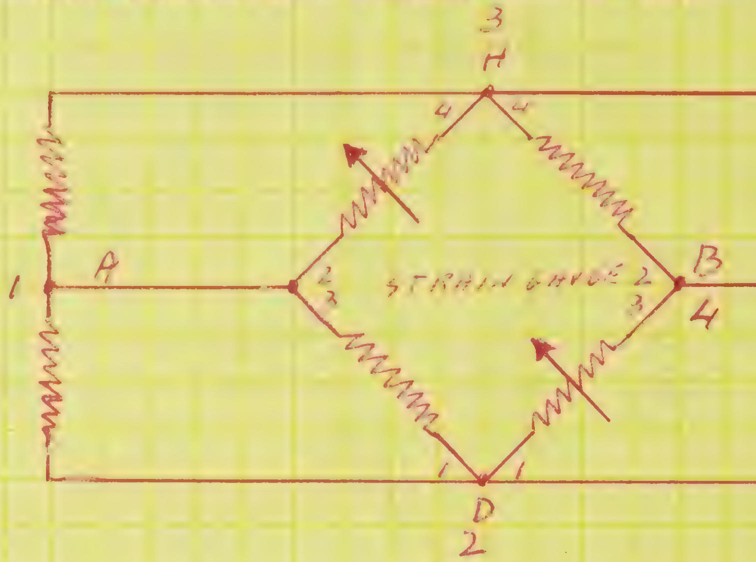
PHASING NETWORK USED TO ELIMINATE PHASE SHIFT IN THE KLUDGES'





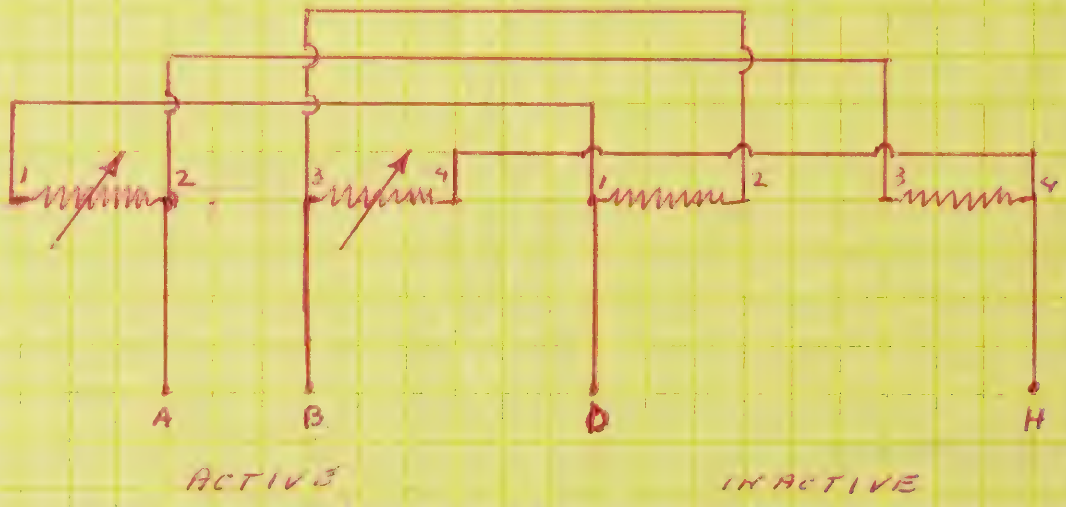
STRAIN GAUGE CLUDGE

SIMULATED STRAIN GAUGE BRIDGE



BALANCE COMPONENTS (R or C) ACROSS $\underset{1-3}{A-H}$ OR $\underset{1-2}{A-D}$

GAUGE PLATE



I

STRAIN GAUGE OSC. SETUP PROCEDURE

1 MATCH FILTER TO OSC. CHA.

2 CHECK FOR CORRECT VOLTAGES OSC-BT 150
" FIL = 6.3
AMP BT = 6.7

3 REMOVE ALL CONNECTIONS FROM CLUDGE.

REPLACE R15 [illegible] WITH 31Ω

4 INSERT OSC. AND SEE IF IT WORKS

5 CHECK OSC. OUTPUT TRANS. WITH A PAIR OF 180Ω RES. ACROSS PINS D & H. MEASURE A.C VOLTA A THE JUNCTION OF THE 180Ω. WITH A VTVM HAVING ONE SIDE GROUNDED. THE VOLTAGE SHOUL BE LESS THAN .01 VOLTS. REVERSING THE 180 RESISTORS SHOULD GIVE THE SAME READING. IF A DIFFERENT READING IS OBTAINED THE TRANS FORMER MIGHT BE BAD.

6 HOOK UP BRIDGE AND SHORT INPUT OF OSCILL. TO GROUND. PINS V & B

2 PUT A 10K & 10Ω FROM D OR H. TO GROUND WI THE 10Ω RES. AT GROUND END

7 PLACE PROBE FROM KEITHLEY AMP. AT JUNCTION OF 10K & 10Ω

9 BY VARYING THE POT ON THE PHASING CKT AT THE HORIZONTAL INPUT OF THE SCOPE OBTAIN 0 PHAS SHIFT ON THE CLUDGE AND WIRING BY GETTING A LINE AT A 45° OR 225° ANGLE OR 135° OR 315° ON THE ~~[illegible]~~ FACE OF THE SCOPE.

10 BALANCING OF THE BRIDGE: GROUND INPUT OF OSCILL. CONNECT KEITHLEY AMP. PROBE TO POINTS A & B IN CLUDGE (ONE AT THE TIME) AND BALAN BRIDGE WITH RESISTANCE NEEDED ON EACH LEG ~~OF~~
* THE BRIDGE WILL BE BALANCED WHEN A PERFECT CIRCLE IS OBTAINED ON THE SCOPE. ~~[illegible] MUST BE [illegible]~~ (THIS INDICATES 90° PHASE SHIFT)

11 INSERT TEST AMPLIFIER AND PLACE TEST PROBE FROM KEITHLEY ON PIN "E" AND READJUST BALANCE COMPONENTS FOR A PERFECT CIRCLE AGAIN. PLACE A SMALL CAP. ACRESS LEGS OF BRIDGE THA GIVE A STRAIGHT HORIZONTAL LINE ON SCOPE. THIS WILL INDICATE A PERFECT BALANCING OF THE BRIDGE.

II

11 CHECK FREQ! REMOVE TEST AMP. AND BY-PASS PLUG BY JUMPING FROM "B" TO "B" AND "A" TO "J". RE-MOVE GROUND SHORTS AND CHECK FREQ. CHANGE THERE SHOULD BE NO CHANGE OR A VERY SMALL CHANGE. THE SAME THING SHOULD ~~THAT~~ BE WITH THE OUTPUT VOLTAGE.

13 OSC. MUST BE SET UP FOR ± 1000 μμ in/in. BY UNBALANCING BRIDGE ACTIVE LEGS (PINS 1-2 & 3-4) WITH 175K FOR HIGH END OF BAND AND PINS 2-4 & 1-3 FOR LOW END. ADJUST R-15 IN OSCI. FOR PROPER SENSITIVITY. 85% B.W.

14 REMOVE JUMPERS. GROUND PIN V. INSTALL AMP. THAT CAME WITH OSC. TO BE SET UP FOR 60 μμ in/in BY UNBALANCING BRIDGE WITH 2.196 M.

A SET BIAS AT EMITER OF Q_2 TO READ 2.45 V. BY INSTALLING A 500K POT FOR R_2 TO FIND THE RIGHT VALUE OF RESISTANCE.

B. INSTALL A BALL PARK VALUE OF 33K FOR R_8 IMPEDENCE MATCHING RES.

C INSTALL A 1K. POT FOR R_6 - GAIN RESIS. - TO FIND THE RIGH VALUE OF RESIST. MAKE SURE AMP. BT IS 6.8 V. D.C.

D. TO MEASURE BIAS AT "E" OF Q2 USE VTVM. AND PUT STRAIN ON THE BRIDGE (2.196 M) TO SEE IF OSC MODULATES. (TWO POINTS IN POT WIL GIVE THE RIGHT BIAS, BUT ONLY ONE IS RIGHT, THE ONE THAT GIVES MODULATION).

E. ADJUST R_6 FOR PROPER SENSITIVITY (85% BW) CHECK OUTPUT VOLTAGE. DIFFERENCE FROM CAL & TO NORM. IT SHOULD NOT VARY MORE THAN ±.2 V. TO ADJUST CHAGE PHASING CAP. IN OSCILL.

F. SHORT 100 Ω R. INSTALL A BALL PARK VALUE OF 33K FOR R8 (IMPEDANCE). READJUSTING GAIN RESISTOR FOR 85% BW. AND BIAS RES. FOR 2.45 V. TAKE A FREQ READING. RECORD

G MOVE SIMULATED STRAIN RES. TO OTHER SIDE OF BRIDGE AND TAKE ANOTHER FREQ REA-DING. TAKE DIFFERENCE OF BOTH REA-DINGS.

H PLACE 500 Ω RESISTORS IN SERIES WITH OUTPUT OF BRIDGE TO AMP. AND ~~NOT~~ TAKE FREQ READINGS AT BOTH SIDES OF BRIDGE, (Ft & F [illegible] SUBTRACT LOW FROM HIGH. TAKE [illegible]S ~~DIFFERENCE~~ THE RATIO OF FREQ. CHANGES ON BOTH STEPS (500 Ω IN AND 500 Ω OUT) THIS RATIO SHOULD BE FROM 2.04 TO 2.08. IF THIS FIGURE IS NOT OBTAINED, REPLACE R8 AND GO THROUGH STEPS F, G, H AGAIN; UNTILL THE PROPER RATIO IS OBTAINED.. THIS INDICATES A PROPER IMPEDENCE MATCH OF 600 Ω. R6-VALUE SHOULD BE FROM 30-60 K

I ~~INSTALL AMP COVER AND TAKE "Z" READING AGAIN.~~

INSTALL R2 FIRST AND RECHECK BIAS VALU AFTER SHORT COOLING PERIOD.

NEXT INSTALL R8 AND CHECK FOR "Z"

" " R6 " " FOR 85% BW

INSTALL AMP COVER AND TAKE "Z" READINGS IF THE OSC. PHASING IS WRONG FIX IT NOW C-15 WILL AFFECT "Z" MEASUREMENTS.

15 SET UP OSC FOR DESIRED RANGE. CHECK IN CHAR FOR RIGH VALUE OF RESISTANCE NEEDED TO SIMU-LATE STRAIN. FOR Ft, Fc & F—.

A SET UP FOR Fc; IF NOT 0 STRAIN, PLACE NECESARY STRAIN ON BRIDGE AND BRING FREQ UP OR DOWN TO CORRESPOND WITH CAL 0 FREQ BY BALANCING BRIDGE WITH RES. (LEGS 1-2 UP) (LEGS 1-3 DOWN) ALSO INSTALL BALANCING CAP. IF NEEDED FOR VOLTAGE.

B SET OSC & AMP. FOR Ft & F— BY USING RES. FOR PROPER STRAIN. IF SENSITIVITY IS TOO HIGH. INSTALL THE PROPER SHUNT RES. ACROSS A & B OF AMP TO BRING THE OSC. SWING TO 85% BW.

IV

16 SET CALIBRATES AND BALANCE OUTPUT VOLTA
GE, BY PLACING R_C & C_C, OR C/R_C

17 RECORD VALUES OF R_C, C_C, BALANCE COMPONE
- SHUNT ACROSS "A" & "B", COMPONENTS INSIDE OSC.
C_2 - C_4 - C_{15}, R_{15}, R_5, R_6, COMPONENTS IN AMP.
R_2 - R_6 - R_8.

HAVE UNITS ~~ALL~~ INSPECTED AND SEN
AMP TO PET.

AFTER BRIDGE HAS BEEN BALANCED AND PHASED PROPERLY, WITH AMP. IN AND INPUT TO OSC SHORTED; CHECK FREQ. THROWING COMMON MODE SWITCH TO POSITION 1 & 3 SHOULD NOT CHANGE THE FREQUENCY TO A GREAT EXTENT; A FEW CYCLES IS O.K, THERE SHOULD BE 180° PHASE SHIFT ON SCOPE BETWEEN POSITIONS 1-3 BUT NOT A FREQ. CHANGE, IF IT IS THE BRIDGE IS NOT PROPERLY BALANCED.

RESISTANCES FOR SIMULATOR STRAIN

25 micro in/in		=	7 M.
50	"		3.5 M.
~~75~~	~~"~~		~~2.332 M~~
60	"		2.916 ~~2.[illegible]~~ M.
75	"		2.332 M.
100	"		1.75 M.
120	"		1.458 M.
125	"		1.4 M.
150	"		1.166 M.
200	"		875 K.
250	"		700 K
300	"		583.3 K
350	"		466.1 K
400	"		437.5 K
500	"		350 K
600	"		291.6 K
1000	"		175 K
1200	"		145.8 K
3000			58.333 K
450	"		388.888 K
1500	"		116.666 K
700	"		250 K

FORMULA USED

$$R_C = \frac{R_O}{S\,(GF)}$$

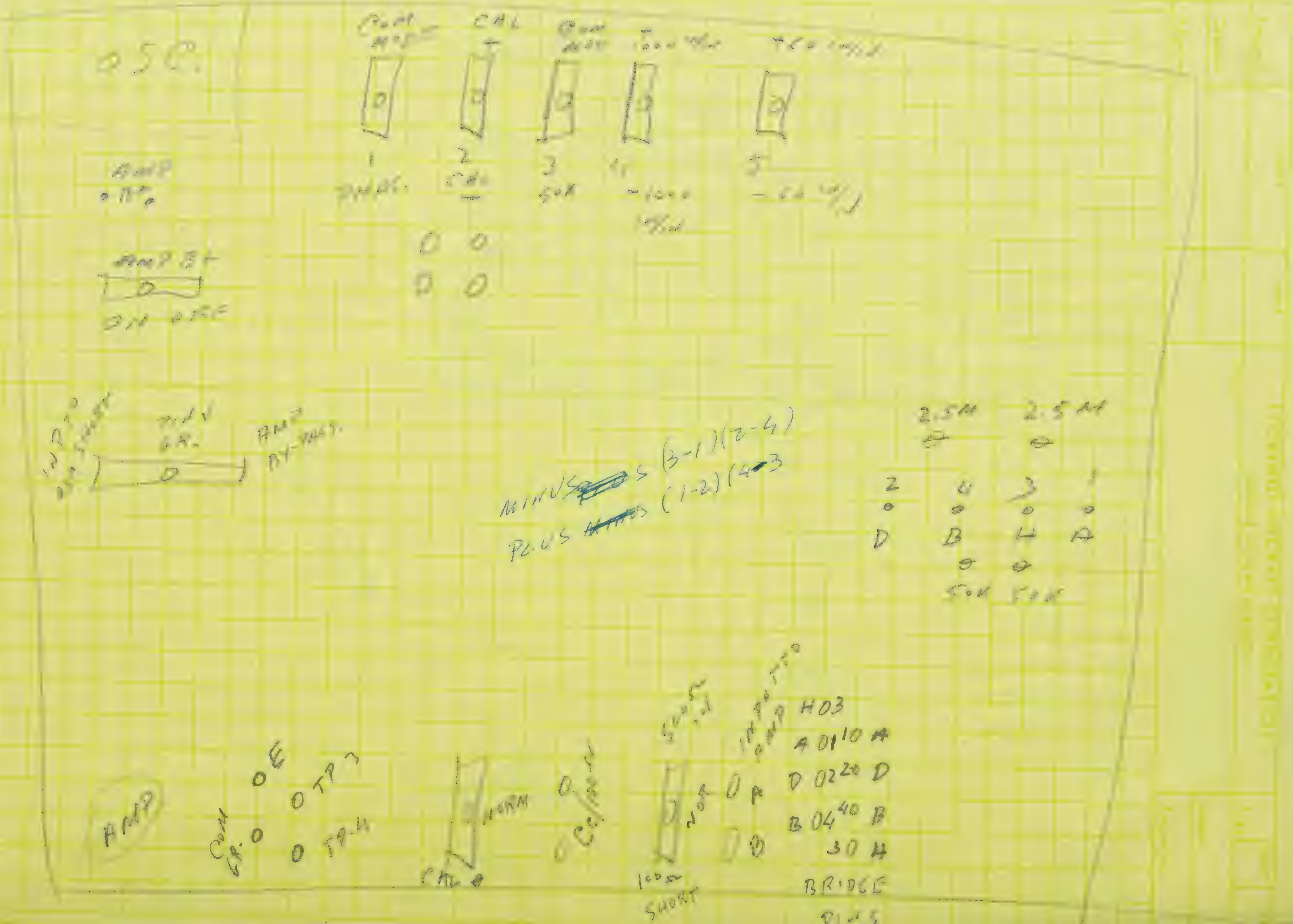
WHERE:

R_C = RESISTANCE NEEDE
TO SIMULATE STRA

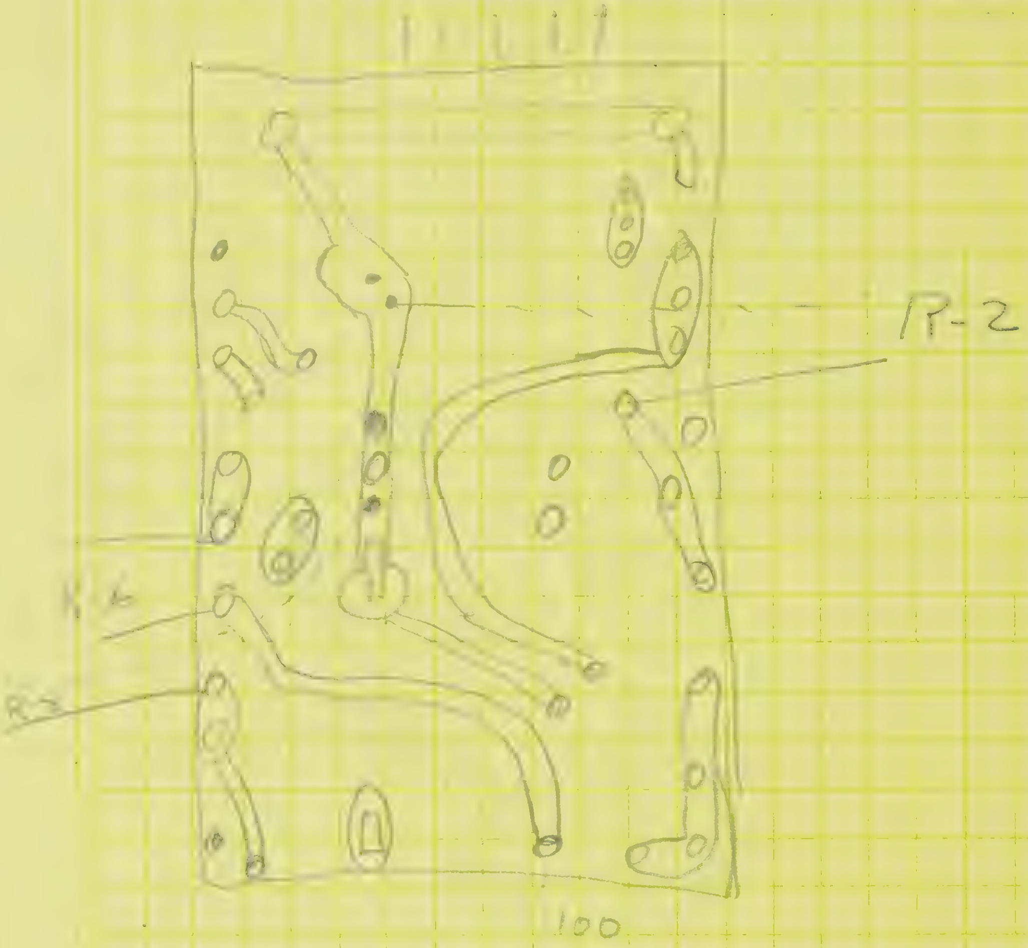
R_O = GAUGE RESISTANCE
IN THIS CASE = 350

S = STRAIN TO BE
SIMULATED

GF = GAUGE FACTOR = 2


OSC.
COM AMP
CAL +
COM AMP
- 1000 M
+ 50 M
1
2
3
4
5
PHASE
CAL -
50K
- 1000 M
- 50 M
AMP
AMP B+
ON OFF
INPUT SHORT
PULL GR.
AMP BY-PASS
MINUS (3-1)(2-4)
PLUS (1-2)(4-3)
2.5M
2.5M
2
4
3
1
D
B
H
A
50K
50K
AMP
COM GR.
E
TP 3
TP 4
CAL
NORM
CAL
500 Ω
NOR
100 Ω SHORT
INPUT AMP
A
B
H03
A 01 10 A
D 02 20 D
B 04 40 B
30 H
BRIDGE

a
a
h
a
p


R-2
K6
R3
100

LOCKHEED

MISSILE SYSTEMS DIVISION
SUNNYVALE, CALIFORNIA

GUS LUCAS
5137 GRANADA STREET
LOS ANGELES 42, CALIF.

LMSD 450613
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MISSILE SYSTEMS DIVISION
XN NAVY TRAINING SCHOOL

Contract NOp-834
Course Notes SC-1-0

TOPIC: INTRODUCTION TO TELEMETRY

1. General

A development program for a missile such as the Navy Polaris requires specialized instrumentation capable of providing in-flight missile performance data. A wide variety of information must be monitored including aerodynamic measurements, control attitudes, guidance parameters, space coordinates, acceleration vibration, pressure, strain, critical operating voltages and temperature. Between 200 and 300 separate inputs may be monitored on a given test flight. These measurements demonstrate the functional reliability of the missile in an operational environment and provide the design engineer with vital data on the performance of the guidance, air-frame and propulsion systems.

The link between these measurements and the data reduction facilities on the ground is called a radio telemeter. It accepts inputs and transmits them over three separate radio frequency links. This telemetered data is almost the only justification for the cost of developing and testing flight vehicles. It has been estimated that 85% of the tangible results of a missile test flight are contained in the telemetry records. Therefore, it is essential that the telemeter system be reliable and that the data be accurate.

ing space. If the primary flight objective is re-entry body development, however, over 90% of the total transducers will be located in the re-entry body, and 80% of these will be temperature pickups.

2. Development Program

To meet the pressing requirements of the Polaris Flight Test Program, Lockheed Missile Systems Division initiated a transistorization program for the Polaris Telemeter. Lockheed was able to draw on 10 years of telemetry development experience, which achieved accuracies of 1% and 2%, on a routine basis, in such high performance vehicles as the X-17, the X-7, the Q5 and the early Polaris-type test vehicles.

Due to the accelerated flight test program, early test flights used whatever equipment was available. The development program, however, has achieved dramatic reduction in size and weight. By eliminating the need for filament supplies, transistorization has also achieved appreciable reductions in power requirements. Lower operating voltages have eliminated arcing and have eased pressurization problems in high altitude operation. The use of etched boards, compact construction, and short leads, naturally improves ruggedness and reliability, especially in high shock environments. Transistors are also much less susceptible to vibration than are vacuum tubes.

System flexibility has been realized through the use of Modular construction throughout the three telemetering links. Phase Sensitive Converters, power supply monitors, subcarrier oscillators and DC amplifiers are all packaged in standard plug-in modules. Changing flight test objectives can be met by selecting and plugging in the required modules. The different links have slots for up to 32 different modules. The basic wiring is the same for the telemeter system in each bird. What wiring modifications are required can be made by simply jumpering connections on a Modification Terminal Board provided for the purpose. Thus, a "shelf-item" system capability is realized despite the change in test objectives with each flight.

Standardization of modules results in important savings in the manufacturing process, but even more important, it increases the performance and reliability of each unit. By not having to design special instrumentation for each test flight, the development engineer can concentrate on improving the design of the standard modules.

3. Qualification program

Because of the accelerated flight test program, there is little time for flight qualifying the new transistorized telemeter system. Some components have flown as passenger items on early Polaris test flights, but basically the confidence level in each unit has been achieved by putting each prototype component through an exhaustive qualification program.

Each unit is first given an electrical test. Its drift characteristics, stability, linearity, sensitivity and output under varying conditions are thoroughly investigated. It must meet stiff specifications. Tolerances must be met with standard parts and no selection of transistors is required. Design emphasis is also placed on simplifying the set-up procedures.

Following the completion of the electrical tests the units are put through environmental tests. The Environmental Laboratory tests each unit at temperatures between 20°C and 85° C, at humidities up to 95%, at altitudes up to 200,000 feet. The vibration, shock and acceleration tests exceed anything the unit may be subjected to in the Polaris environment.

Following the environmental tests a careful analysis is made of test results. Any unit failures are pin-pointed as to cause. Close coordination is maintained between design and manufacturing engineers. Some failures may indicate a need for closer control of manufacturing tolerances. Others may be due to faulty mounting or incorrect location of parts. When the qualification tests uncover a flaw which requires circuit redesign, a careful analysis is made so that specifications can be met without resorting to selection of critical parts. Each change or redesign is then tested again and again until the unit successfully passes all phases of the qualification program.

Many of the telemetering modules are also tested on the System Simulator. This test set contains all major components of a complete telemetering link, including inputs, commutators, power sources and transmitter. It will accomodate all plug-in units and enables investigation of the compatibility of the units in the system, and pin points such problems as cross-talk, etc., in advance of system checkout.

After each component is thoroughly "wrung-out", the entire telemeter link is then given a bench check-out. Inputs are supplied to all channels. In some instances transducers can be simulated by electrical voltages. If not, transducers are wired in and calibrated input quantities applied. The output of each channel is monitored and the entire link is tested for accuracy, stability and repeatibility. Finally, the entire telemetering link is subjected to an environmental test.

Final test reports indicate that the transistorized telemeter will meet all specifications and give the same high quality data supplied by the vacuum tube telemeter in previous test flights. It will do this with units that occupy less space, weigh less, and drain only a portion of the power formerly required - making space and power available for other instrumentation which will also play a vital role in the development of the Polaris.

4. FM/FM - Telemetry

The initial type of telemetering equipment to be used in the Polaris AX series of vehicles is the FM/FM system.

The telemetry equipment installed in the vehicle is contained in the telemetering units. Each of these units is semicircular in cross-section, has an approximate radius of 7½ inches, and is 14 inches in length. Units No. 1 and No. 2 are grouped together to form a full cylindrical shape and are installed in the re-entry body of the vehicle. Stand-off bracketry on the periphery of the units permits installation of these two units in the re-entry body compartment just aft of the telemetering antenna. Unit No. 3 is installed as part of an equipment grouping with the AZUSA installation in the equipment section of the vehicle.

The container of each telemetering unit has internal, welded bracketry to accept the various components which are packaged as an assembly to permit easy removal. These containers are also sealed and air-cooled to facilitate a heat dissipation of approximately 200 watts per unit. The unit electrical connectors are potted to prevent condensation and arcing.

Power Supplies for telemetry are external to the telemetering units. These composite power supplies also furnish power for the beacon and destruct installations.

The FM/FM T/M System uses the FM/FM telemetry system because of the volume of information which can be transmitted over the number of continuous and commutated channels. The FM/FM system provides flexibility of use along with several desirable features such as simplicity, ruggedness, small size, and low weight.

In the FM/FM System each variable quantity to be measured is converted into an equivalent electrical signal. The frequency of this signal modulates a subcarrier oscillator operating in one of the standard Research and Development Board (RDB) subcarrier frequency bands. The output voltages of the subcarrier oscillators are combined and applied to the modulator circuit of the VHF radio transmitter which, in turn, excites the transmitter antenna.

The transmitted signal is recovered by the ground station receiver which converts the FM radio frequency signal into a composite subcarrier voltage. Band-pass filters separate the individual subcarrier voltages from the composite signal. The filtered subcarrier signals are fed into subcarrier frequency discriminators which produce a varying DC voltage output. The current output of the discriminator is the electrical equivalent of the original stimulus and may be fed into a galvanometer type recording oscillograph to convert the electrical signal into a permanent photographic record. By use of a calibration chart, or by automatic reduction equipment, the information may be converted into terms of the original quantity for the purpose of analysis.

Various types of pick-ups are used to provide the required data from the vehicle. Among these are thermocouples, resistance bridges, variable resistance transducers, and variable reluctance transducers.

The FM/FM T/M units installed in the vehicle use the following two types of multi-channel systems:

Frequency Division Multiplexing. Frequency division multiplexing provides continuous information. However, within each telemetering unit, it is limited to 18 samplings by the 18 specified channel bands. In addition, receiving station limitations generally restrict the number of usable bands to less than 18. In this type of transmission, each measurement is assigned to a different subcarrier frequency band. The individual measured quantity is used to frequency modulate a subcarrier oscillator within the limits of the assigned frequency band. The output signals from the subcarrier oscillators are combined and the composite signal is used to modulate an RF carrier. At the receiving station, the composite subcarrier signal is recovered through the use of a frequency modulation receiver which separates the composite signal into the individual subcarrier signals by the use of bandpass filters. Subcarrier discriminators demodulate the individual subcarriers and produce a DC output which is proportional to the original measured quantity.

Time Division Multiplexing. Time division multiplexing provides repitive, sequential sampling. As indicated above, a limited number of continuous intelligence pick-ups may be combined in a multiplexing technique. If, however, the intelligence sources may be sampled in a recurring sequence, the individual bands may carry a number of measurements sampled in sequence by a commutator switch (TSC). This sampling is accompanied by a substantial reduction in the rate of intelligence change (frequency response) which may be transmitted. These additional complicate the basic receiving elements required.

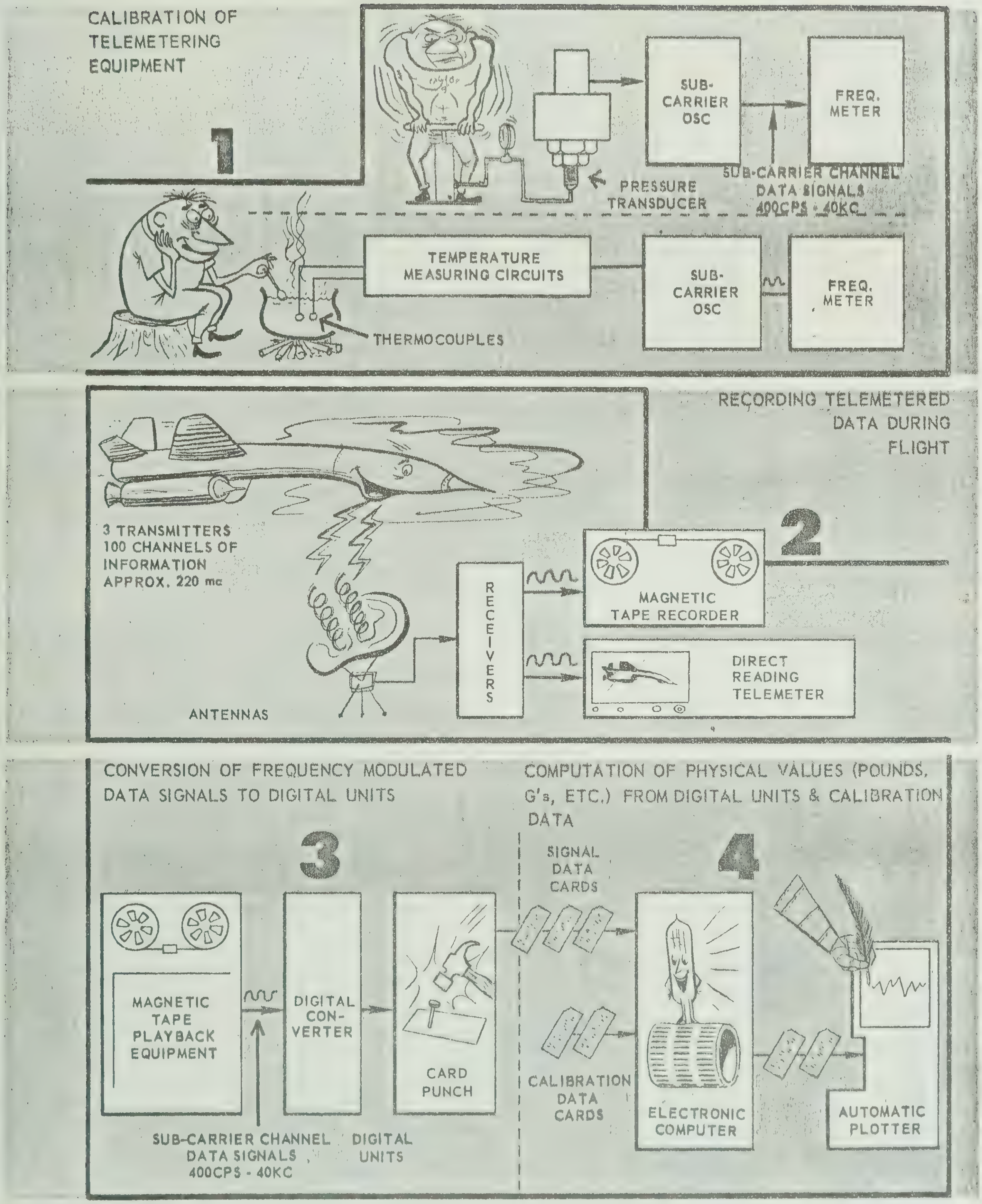
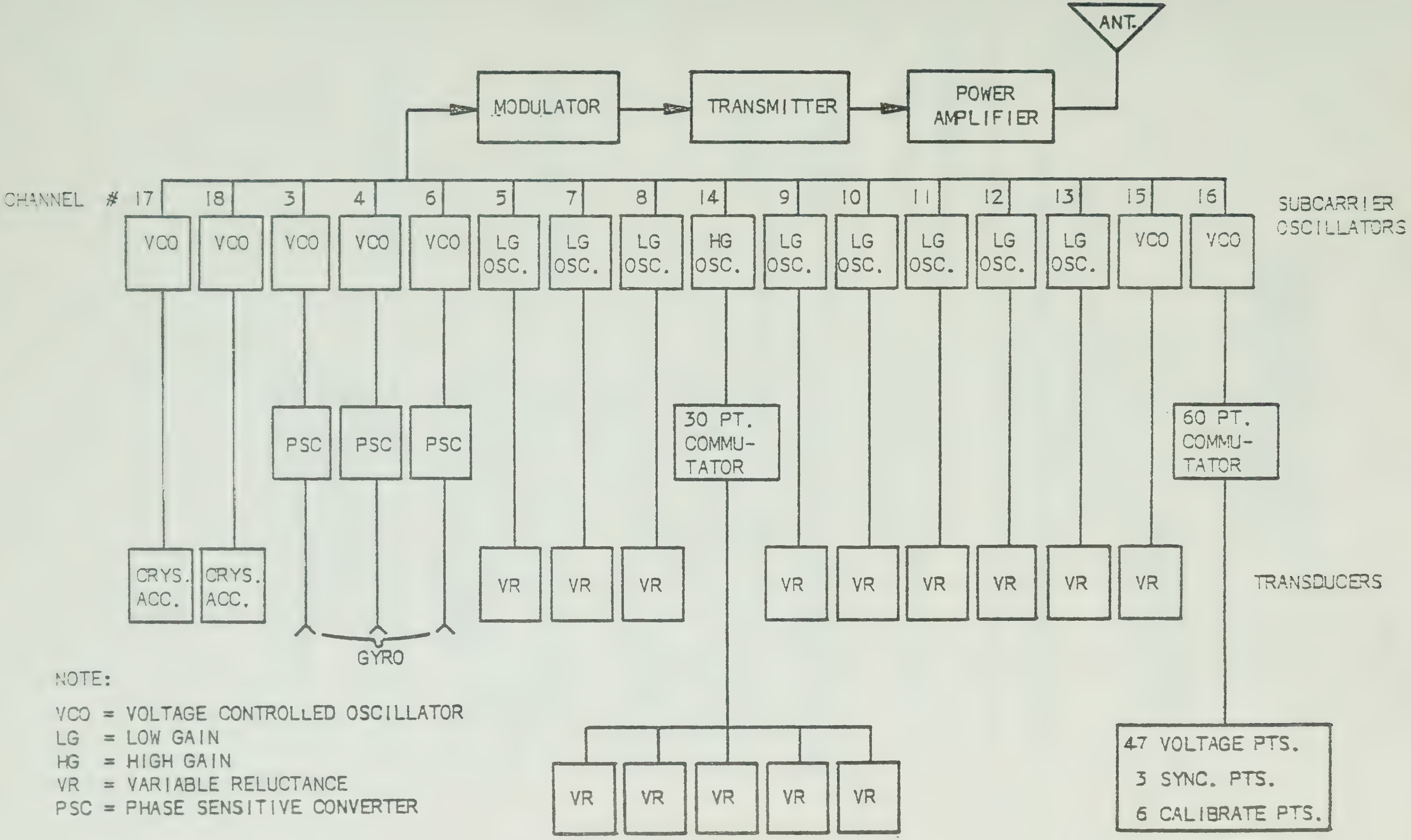

CALIBRATION OF TELEMETERING EQUIPMENT
1
SUB-CARRIER OSC
FREQ. METER
PRESSURE TRANSDUCER
SUB-CARRIER CHANNEL DATA SIGNALS 400CPS - 40KC
TEMPERATURE MEASURING CIRCUITS
SUB-CARRIER OSC
FREQ. METER
THERMOCOUPLES
RECORDING TELEMETERED DATA DURING FLIGHT
2
3 TRANSMITTERS 100 CHANNELS OF INFORMATION APPROX. 220 mc
RECEIVERS
MAGNETIC TAPE RECORDER
DIRECT READING TELEMETER
ANTENNAS
CONVERSION OF FREQUENCY MODULATED DATA SIGNALS TO DIGITAL UNITS
3
MAGNETIC TAPE PLAYBACK EQUIPMENT
DIGITAL CON-VERTER
CARD PUNCH
SUB-CARRIER CHANNEL DATA SIGNALS 400CPS - 40KC
DIGITAL UNITS
COMPUTATION OF PHYSICAL VALUES (POUNDS, G's, ETC.) FROM DIGITAL UNITS & CALIBRATION DATA
4
SIGNAL DATA CARDS
CALIBRATION DATA CARDS
ELECTRONIC COMPUTER
AUTOMATIC PLOTTER

Figure 1. A Highly Simplified View of the Principal Operations Involved in Telemetering



NOTE:

VCO = VOLTAGE CONTROLLED OSCILLATOR
LG = LOW GAIN
HG = HIGH GAIN
VR = VARIABLE RELUCTANCE
PSC = PHASE SENSITIVE CONVERTER

FM/FM Telemetry System - Polaris AX

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TOPIC: Telemeter System Characteristics

The system shall conform to the following characteristics, excluding the individual component requirements which are already specified.

Power Requirements:	28 VDC ± VDC ±10% @ 5.75 amps 6.3 VDC @ 2.25 A
Transmitter Frequency Deviation:	±125 KC maximum for composite signal ±115 KC minimum for composite signal
Amplitude of Subcarrier Signals:	Subcarrier signal amplitude to be adjusted in such a manner that the carrier deviation per channel is in approximate accordance with a 3/2 power taper relative to the subcarrier frequency. A minimum of 3 KC deviation per channel should be maintained.
Subcarrier deviation ratio:	Stimulus frequencies should not exceed IRIG specifications for a deviation ratio of 5. Peak to peak subcarrier deviation should not be less than 85% of bandwidth and not exceed 100% of bandwidth.
Commutation Rates:	Maintain rates consistent with IRIG specifications for commutation samples per channel. Indicated RPS ± 10% Use of standard LMSD commutation scheme.

Cross Modulation: Less than 2% RMS total for any channel with all other channels operating.

System Accuracy: System accuracy shall be ± 1% as determined from subcarrier oscillator input to discriminator output.

The measurement accuracy shall be determined under satisfactory conditions of aforementioned system items.

Water cooling shall control equipment temperature to a maximum change of 50°F from temperature at start of test.

A preflight and a minimum of three inflight calibrate operations shall furnish compensation data for system accuracy determination.

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TOPIC: Subcarrier Oscillators

Two different subcarrier oscillators, the Voltage Control (VCO) and the Dual Loop (DLO), are used in the Polaris Transistorized Telemeter System. By using both oscillators, the telemetering system can be adapted to a wide variety of transducers and maximum system flexibility with optimum component useage is assured.

The VCO shown in figure 3 is basically an RC oscillator and the output frequency changes in direct proportion to an input voltage. The direction of the frequency deviation is a function of the polarity of the input voltage and the VCO can be used with either AC or DC signals. It is thus conveniently used with any pickup whose output is a DC or AC voltage proportional to the physical quantity being monitored. Typical examples of such pickups as shown in the figures 3, 4, 20 and 21, are thermocouples, potentiometers, and strain gages. As both thermocouples and strain gages are low output devices, a DC amplifier (TCA) is required to increase the signal level to approximately 1 volt. A potentiometer, if monitoring gyro attitude, is usually AC excited and the excitation phase shifts each side of center position of the potentiometer. As this phase information must also be transmitted, a phase sensitive converter (PSC) is required.

The output of this converter is a DC voltage whose amplitude is proportional to the amplitude of the AC signal and whose polarity is a function of the phase of the AC signal. Critical supply voltages can also be monitored, as shown in the figure, simply by using a voltage divider. The VCO, however, cannot easily be used with variable reluctance transducers.

The DLO shown in figure 2 supplies its own excitation to the transducer. The transducer is usually connected as part of a bridge circuit, the initial balance of which is upset by the physical input to the transducer. The unbalanced bridge signal is fedback and mixed with the oscillator signal, changing its frequency. The DLO can be used with any transducer which can be connected as part of a bridge circuit. This is a convenient arrangement for strain gages and variable reluctance pickups. The figure also shows a typical connection for a potentiometer. The DLO is not normally used with pickups which generate an AC or DC signal. Typical examples would be piezoelectric vibration pickups and thermocouples.

As the DLO supplies excitation (at the channel frequency) to the transducer, the long leads (and resultant high lead capacity) limit its use to channels below 15. Thus the VCO is used almost exclusively with all types of vibration pickups where the required frequency response demands the use of the higher channels (15 through 18). The low frequency capabilities of both oscillators is approximately the same - DLO's can be used on channel 4, VCO's on channel 3.

Since the transducer is a part of the oscillatory circuit of the DLO, each transducer must be calibrated with that DLO and if either component must be replaced, the entire calibration has to be repeated. This gives the VCO a considerable advantage, especially on commutated channels where a large number of transducers are used with a single oscillator. Setup of the VCO requires only that the sensitivity be adjusted for full bandwidth (deviation) for a given standard input voltage - the output of each transducer is then adjusted (if necessary) so that voltage is not exceeded. If any single transducer proves faulty, it is not necessary to recalibrate the oscillator.

On continuous channels the DLO has some advantages over the VCO, especially with low output devices such as strain gages. First, each VCO would require a DC amplifier, doubling the number of required modules. Secondly, DC amplifiers are subject to drift whereas the internal AC amplifiers in the DLO can be feed-back stabilized.

On commutated channels, however, one TCA-VCO combination serves a large number of pickups and commutation is used whenever the frequency response requirements permit. Furthermore, the calibration points on the commutator make it possible to eliminate any small amplifier drift effects during the data reduction process.

The factors determining the choice of subcarrier oscillator (for a given application) are summarized on the following page. It should be emphasized, however, that this represents only typical useage. While the VCO obviously has the greater versatility, the unique advantages of the DLO may dictate its use in any given application.

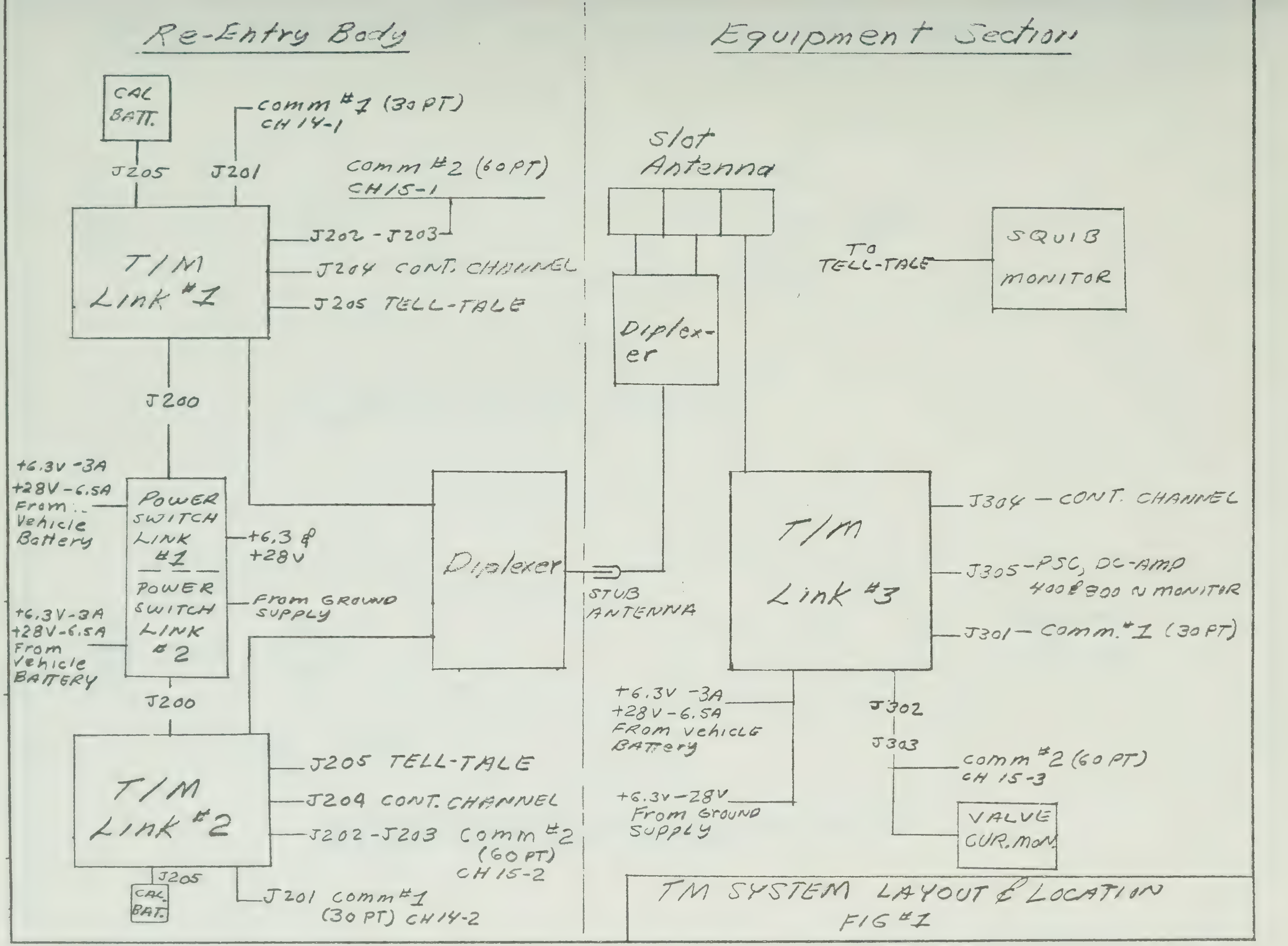
TYPICAL SUBCARRIER OSCILLATOR UTILIZATION

T TO ILLATOR	OSCILLATOR	PICKUP	MEASUREMENT	COMMENTS
C	VCO	None	Power Supply	Setup and calibration of VCO is very simple. DLO requires additional circuitry to monitor voltage.
	VCO	Pot	Actuator Position	Setup and calibration simpler for VCO.
			Acceleration	
			Pressure	
	VCO-TCA	Thermocouple	Temperature	Although extra amplifier is required, one VCO-TCA combination serves many pickups. Thermocouples are primarily low frequency devices and signals can be commutated.
	VCO-TCA	Strain Gage	Acceleration	If frequency response is low, signals can be commutated. Same comments as above paragraph apply.
			Pressure	
C	DLO	Strain Gage	Acceleration	When the frequency response data is desired, the data is sent over a continuous channel. The DLO does not require a separate amplifier for low level inputs.
			Pressure	If frequency response data is required, data is sent over continuous channel. Same comments apply as paragraph above.
	DLO	Variable Reluctance	Pressure	Separate AC generator not required as DLO supplies its own excitation to transducer.
			Acceleration	

TYPICAL SUBCARRIER OSCILLATOR UTILIZATION CONTINUED - -

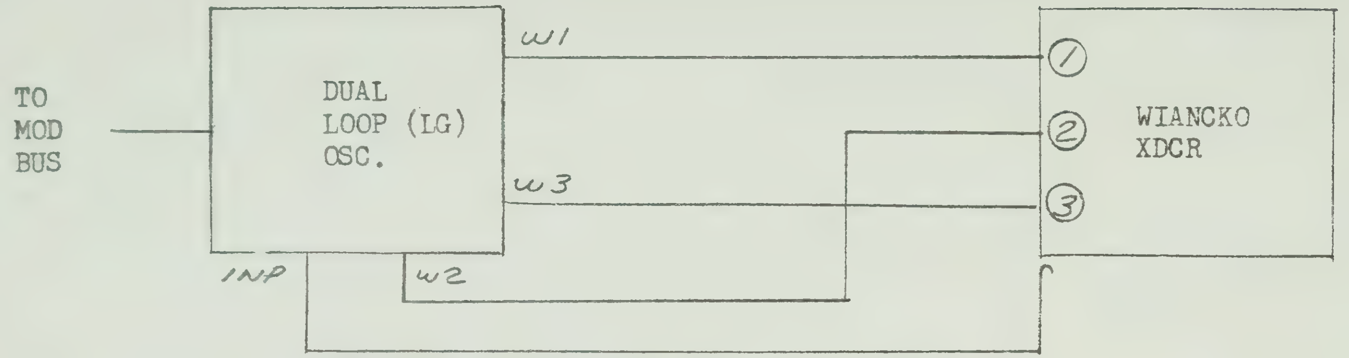
INPUT TO OSCILLATOR	OSCILLATOR	PICKUP	MEASUREMENT	COMMENTS
AC	VCO-PSC	Pot	Gyro, Attitude	Although a PSC is required, the VCO-PSC combination can work directly from signals tapped off the integrated gyro-pickup package. The DLO would require the attachment of an additional pot
		Diff. Trans.	Gyro, rate	
	VCO V.C.**	Piezo elec.	Vibration	High frequency response can only be obtained on channels 15 through 18. DLO will not work with piezoelectric pickups directly, nor can it be used over channel 14.
	VCO-TCA	Strain Gage	Vibration	High frequency response required dictates use of channels 15 through 18.
Amplitude Tell-Tale	DLO or VCO	Break-wire	Event	Either oscillator is equipped for amplitude tell-tale. Event monitored may be first or second stage separation, etc.

** Vibration Coupler

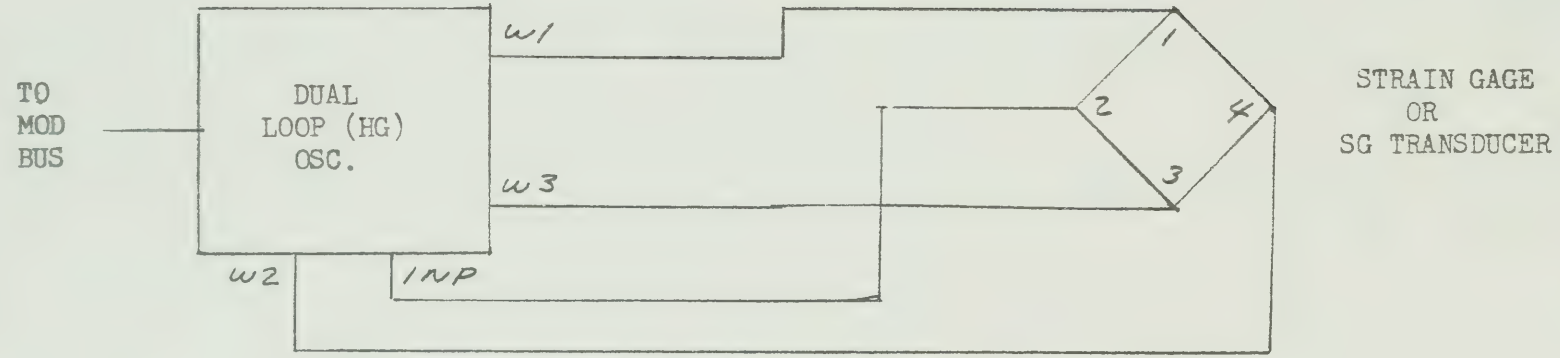


TM SYSTEM LAYOUT & LOCATION
FIG #1

Dual loop oscillator -- continuous channel (dual loop oscillator HG &LG reauires matching with transducer).



TYPICAL CIRCUIT VARIABLE, RELUCTANCE TYPE
TRANSDUCER AND DUAL LOOP LG OSCILLATOR.



TYPICAL SIGNAL CIRCUIT, STRAIN GAGE OR STRAIN GAGE TYPE
TRANSDUCER AND DUAL LOOP HG OSCILLATOR.

TEMPERATURES - Thermocouple type (commutated) and/or Resistance thermometer.

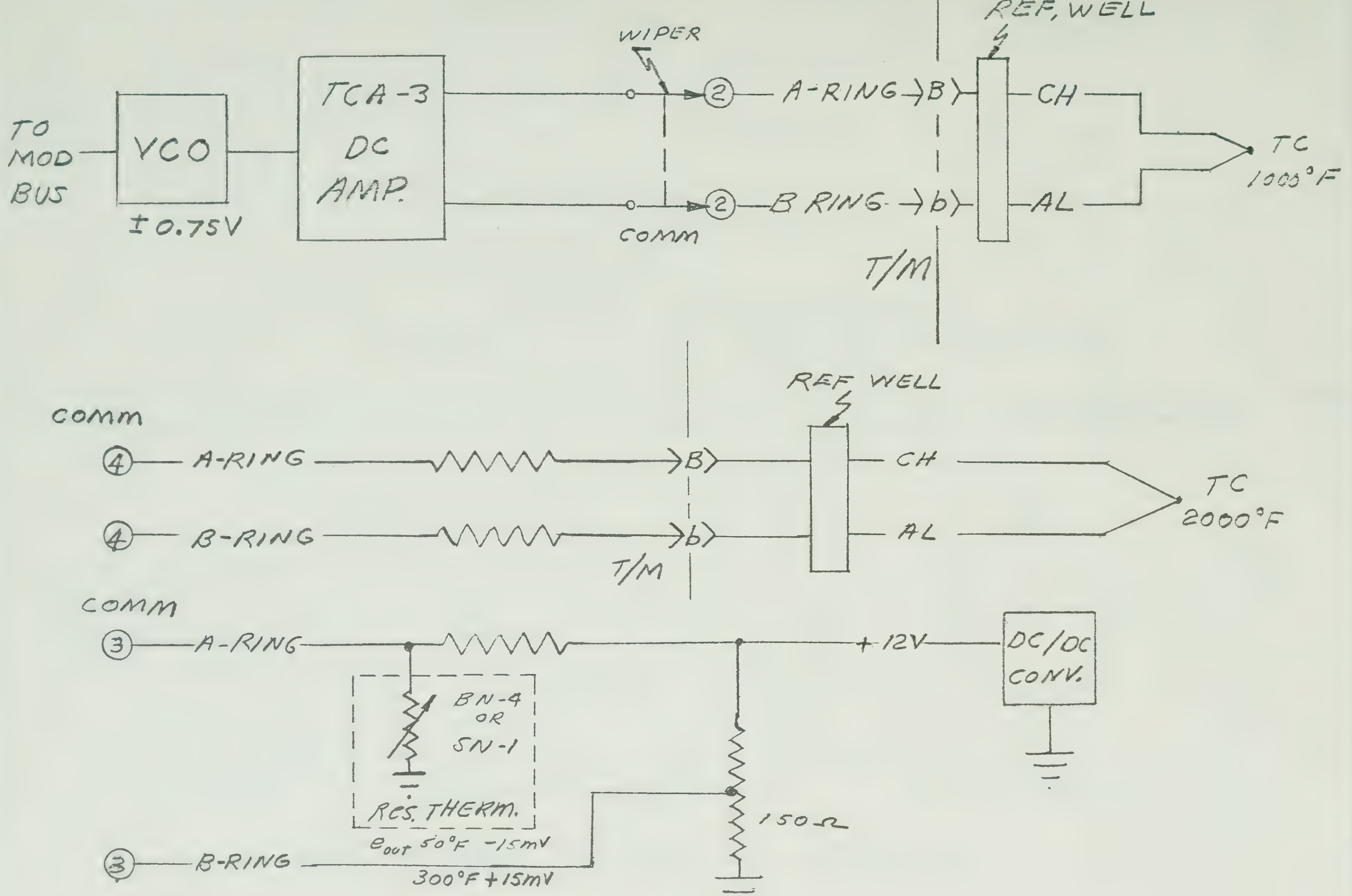
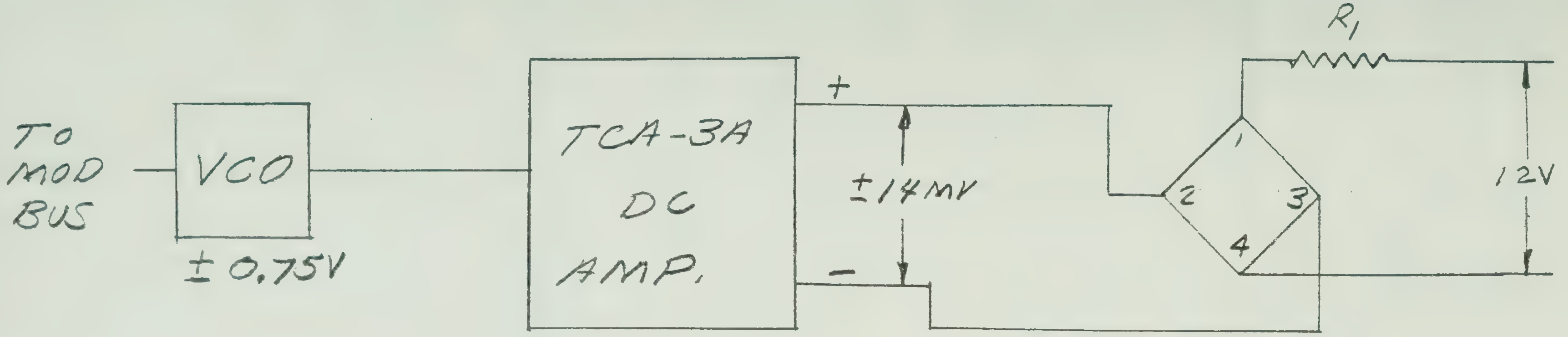


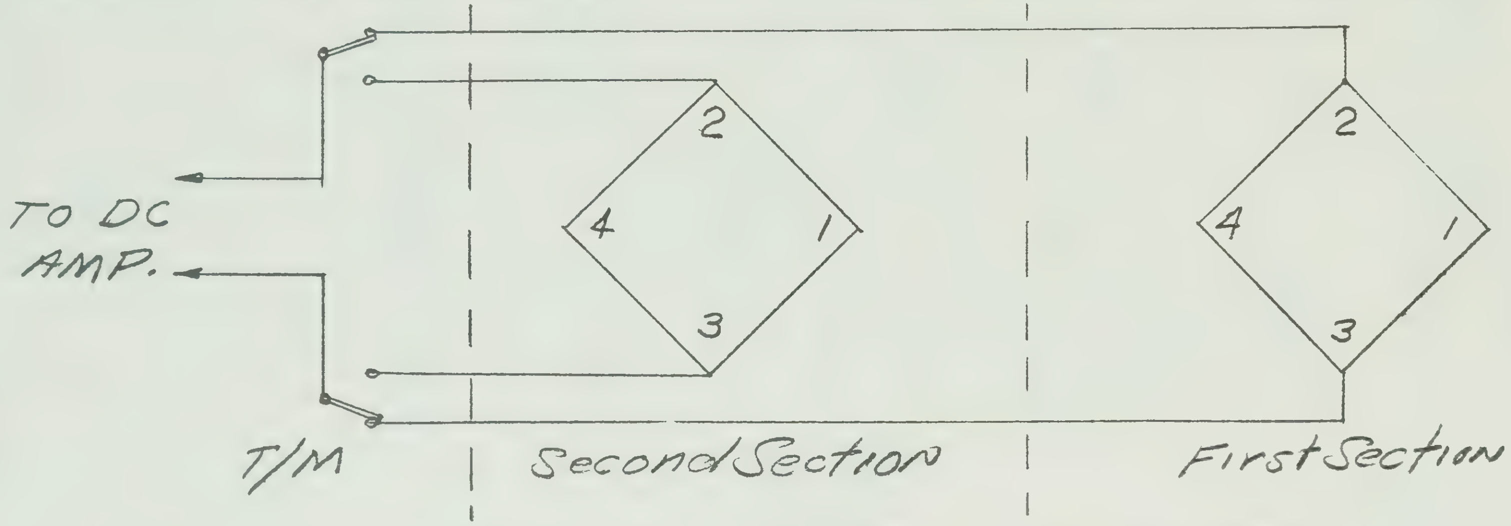
FIG 4

Vibration - strain gage type transducer, continuous Channel



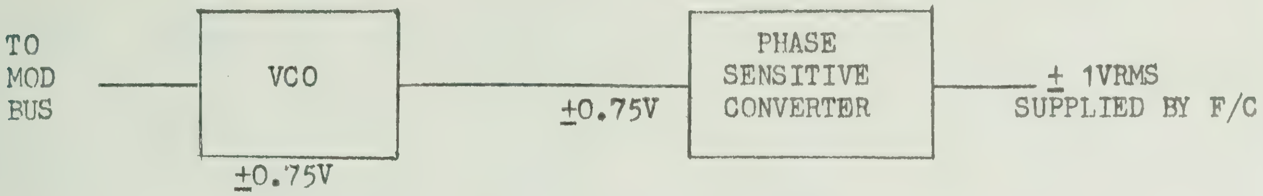
Typical Signal Circuit
Vibration-VCO & SG Type Transducer
FIG. 5

VIBRATION-Strain Gage type transducer, Continuous Channel



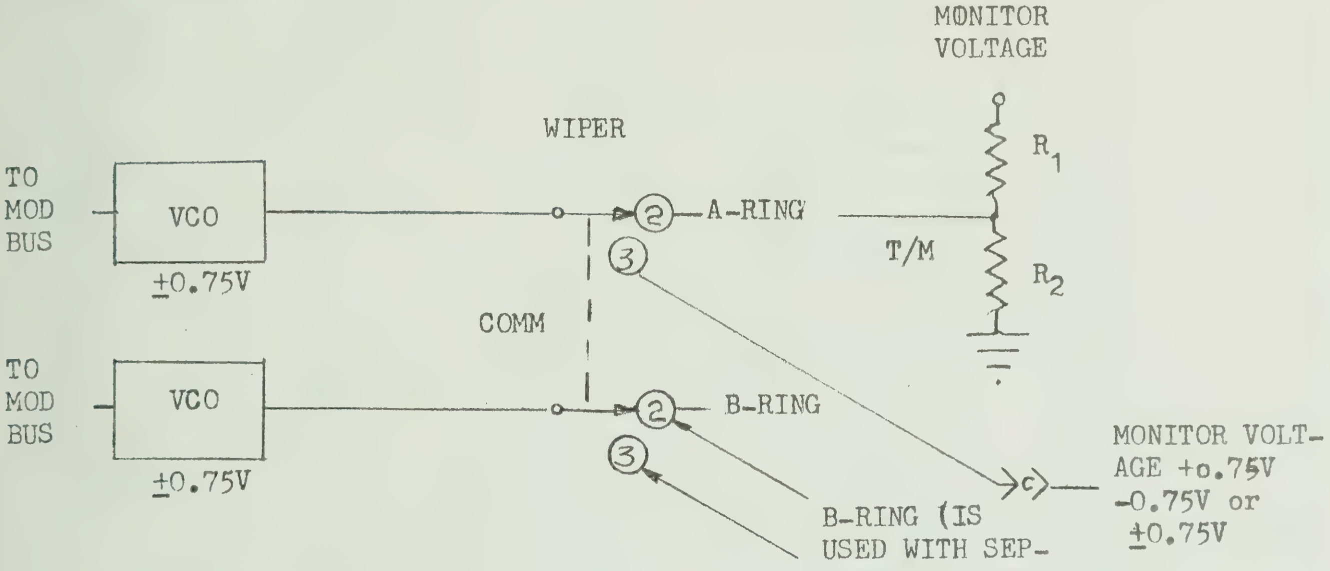
INTERSTAGE SWITCH CIRCUIT
FIG. 6

Flight Control Gyro, accelerometer and amplifier output is 1V RMS 800 CPS in phase or out of phase with reference voltage.



TYPICAL SIGNAL CIRCUIT FIG.7

Voltage Commutated.


MONITOR VOLTAGE
WIPER
R1
TO MOD BUS
VCO
±0.75V
2
A-RING
3
T/M
R2
COMM
TO MOD BUS
VCO
±0.75V
2
B-RING
3
MONITOR VOLTAGE +0.75V -0.75V or ±0.75V
B-RING (IS USED WITH SEPARATE VCO IF REQUIRED)

TYPICAL SIGNAL CIRCUIT

FIG 8

Voltage Controlled Oscillator continuous channel.

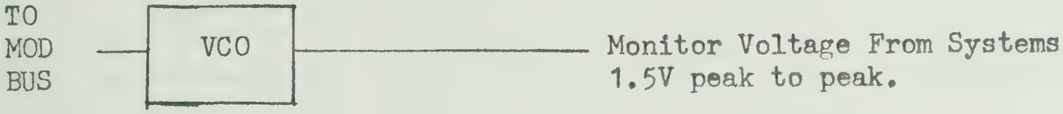
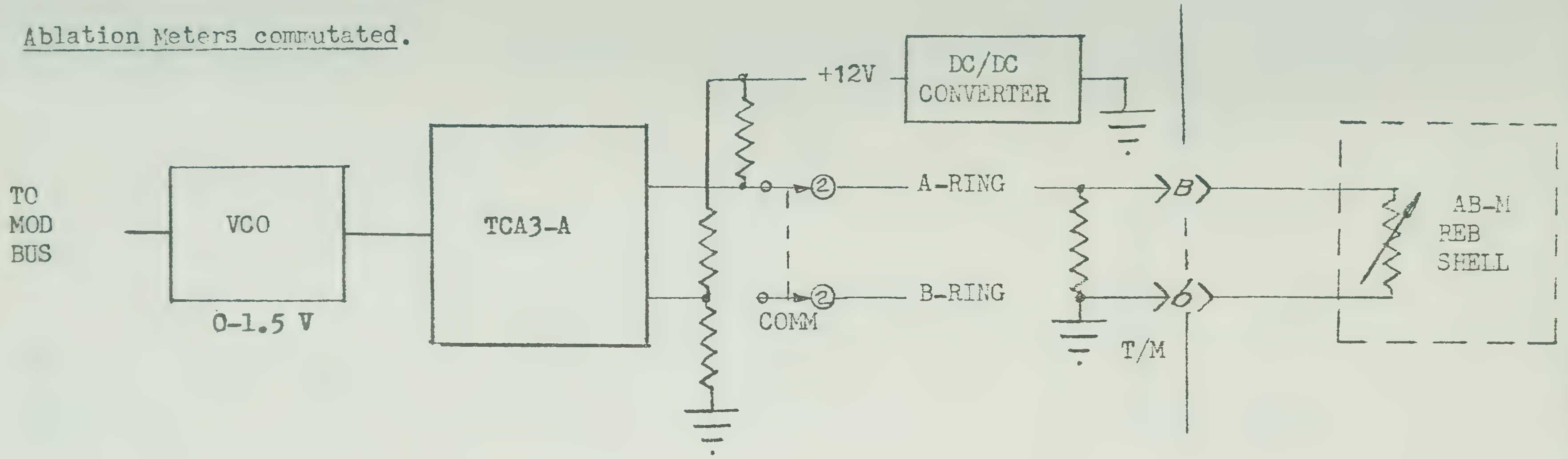


FIG 9

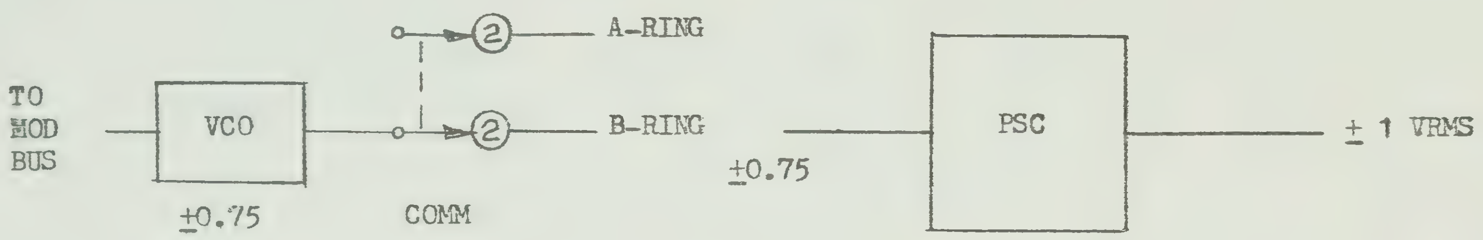
Ablation Meters commutated.



TYPICAL SIGNAL CIRCUIT ABLATION METER

FIG 10

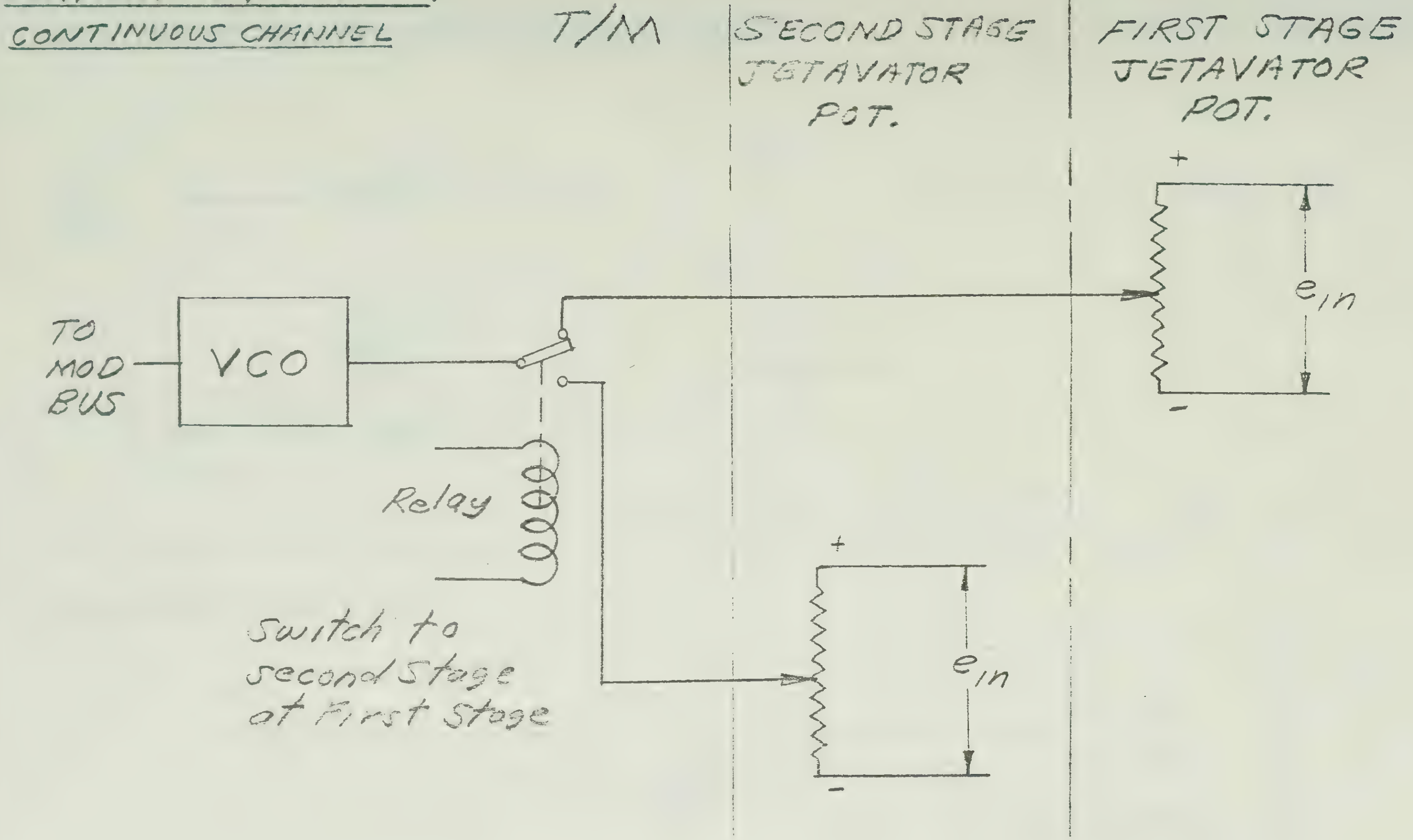
PSC COMMUTATED



TYPICAL SIGNAL CIRCUIT PSC COMMUTATOR INPUTS

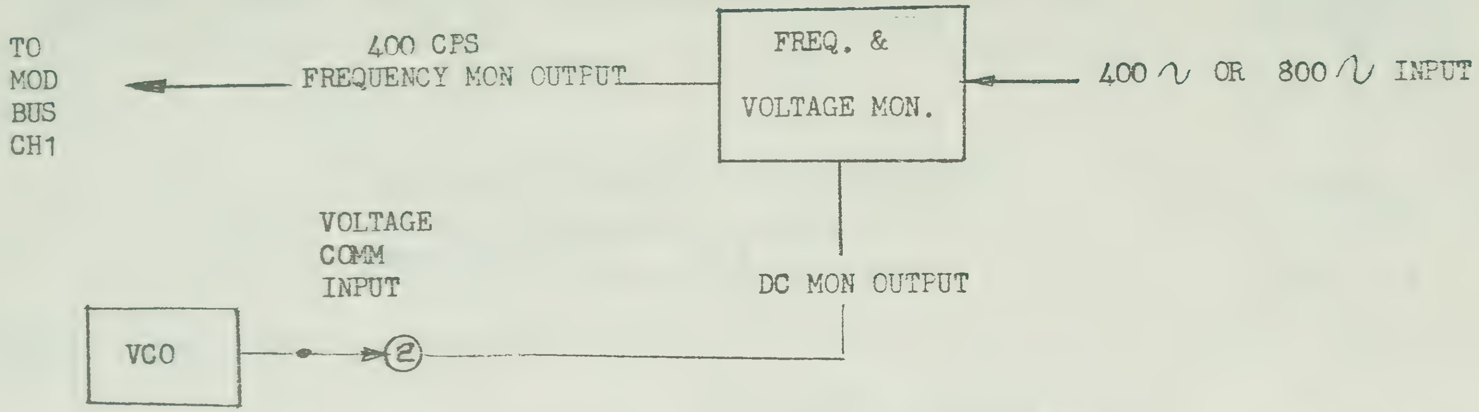
FIG 11

JETAVATOR POSITION, CONTINUOUS CHANNEL


T/M
SECOND STAGE JETAVATOR POT.
FIRST STAGE JETAVATOR POT.
TO MOD BUS
VCO
Relay
Switch to second Stage at First Stage
+
-
e_{in}
+
-
e_{in}

TYPICAL SIGNAL CIRCUIT
JETAVATOR POT.
(FLIGHT CONTROLS)
FIG 12

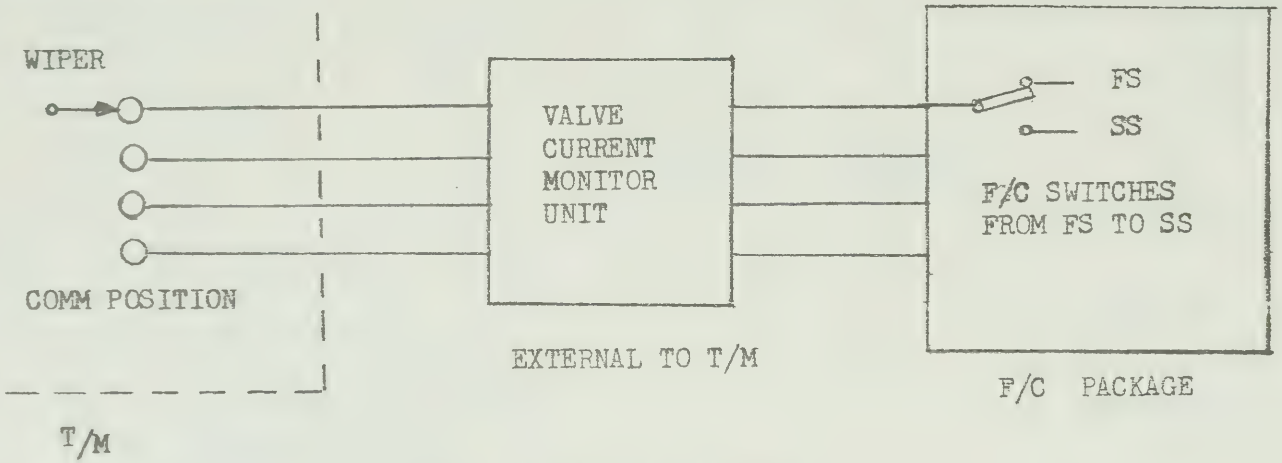
400 or 800 CPS FREQUENCY AND VOLTAGE MONITOR UNIT.



TYPICAL SIGNAL CIRCUIT FREQ. & VOLTAGE MONITOR

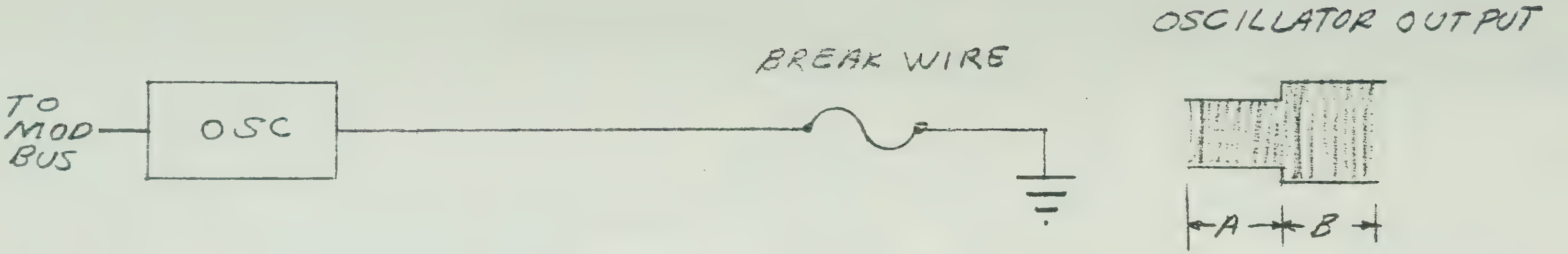
FIG 13

VALVE CURRENT MONITOR UNIT



TYPICAL SIGNAL CIRCUIT

FIG 14

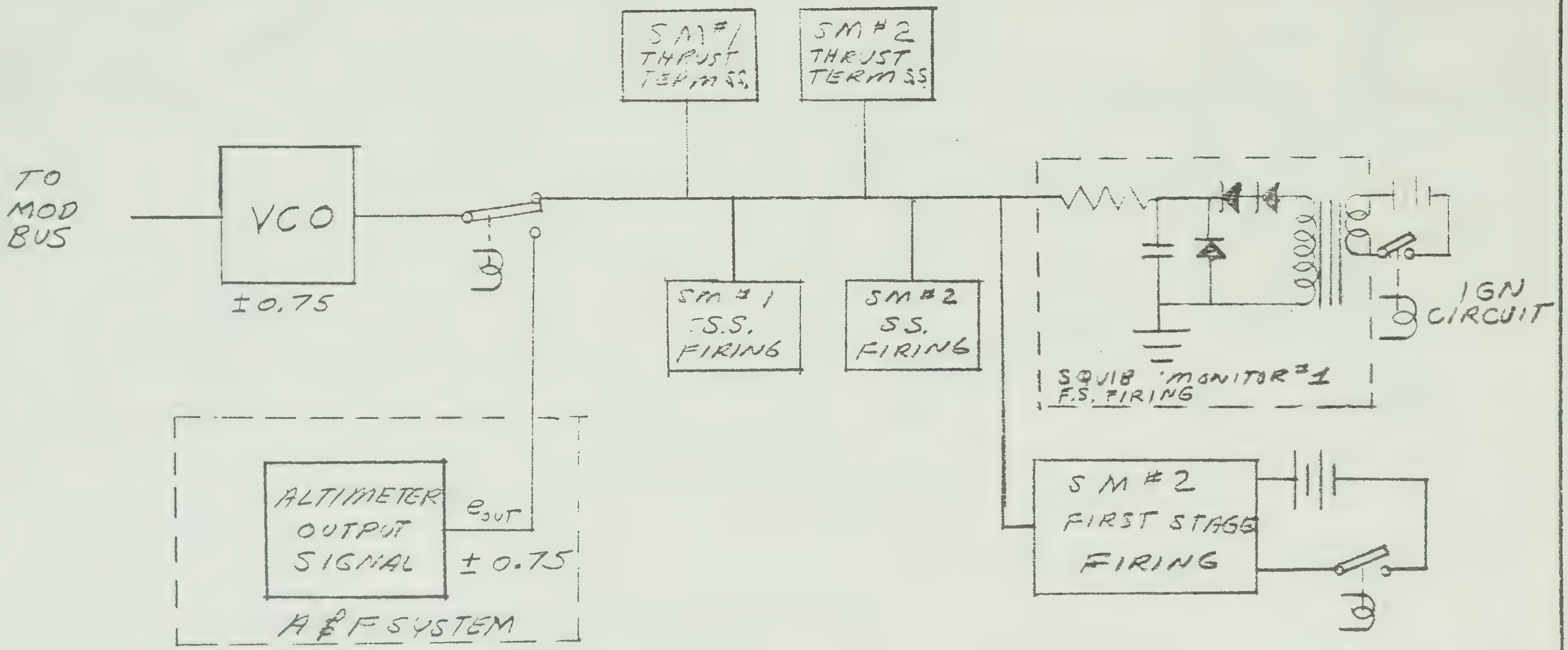


"AMPLITUDE" TELL-TALE

FIRST STAGE Separation
SECOND " "
THRUST TERMINATION-etc.

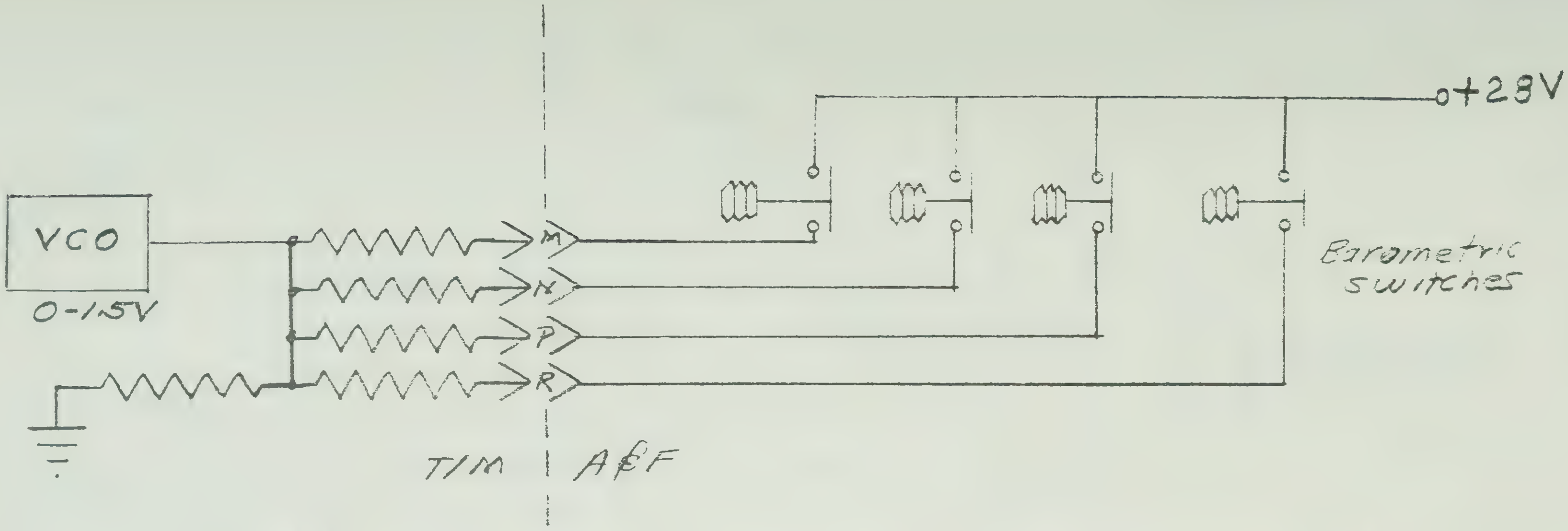
A-Before wire break.
B-AFTER " " .

FIG 1.5



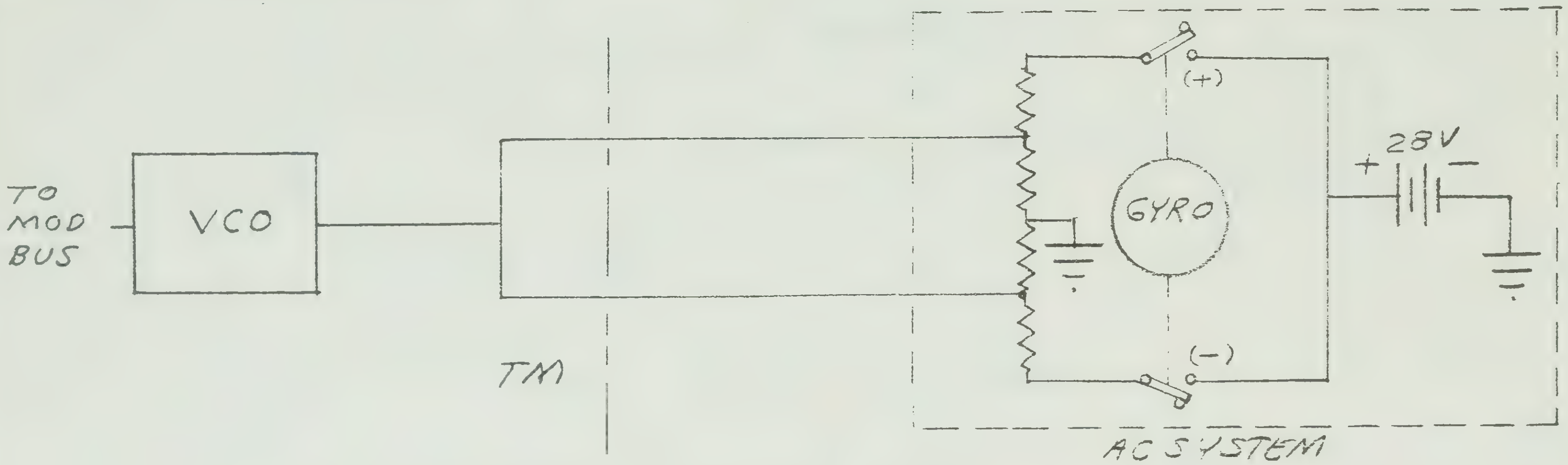
ALTEMETER TELL-TALE

SQUIB MONITORS
FIG 1.6



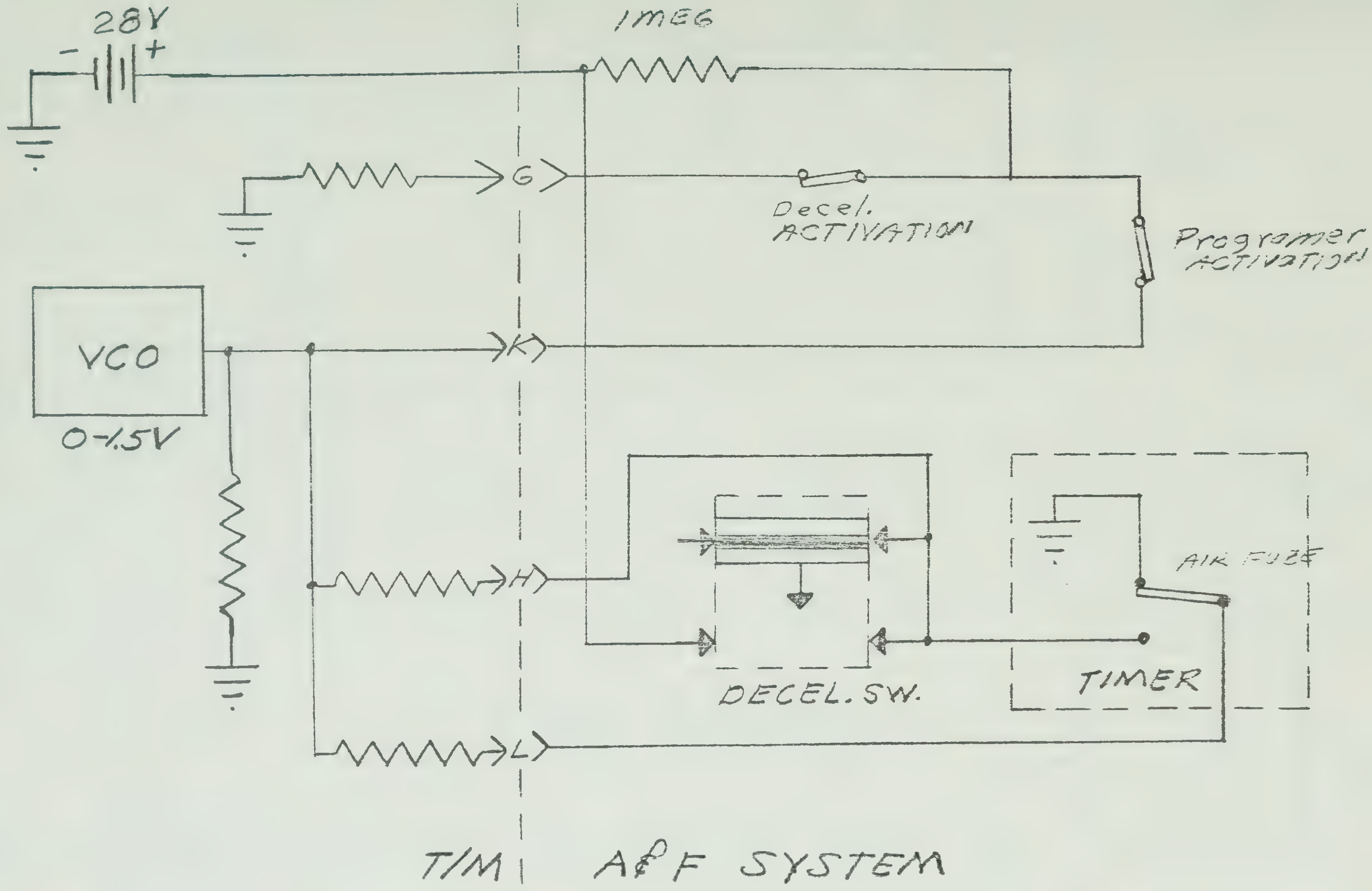
A&F Tell-Tale SIGNAL PATH

FIG 17



GYRO RATE SIGNAL PATH
TELL-TALE

FIG 18



SIGNAL CIRCUIT
TELL-TALE

FIG-19

HYDRAULIC SYSTEM PRESSURE

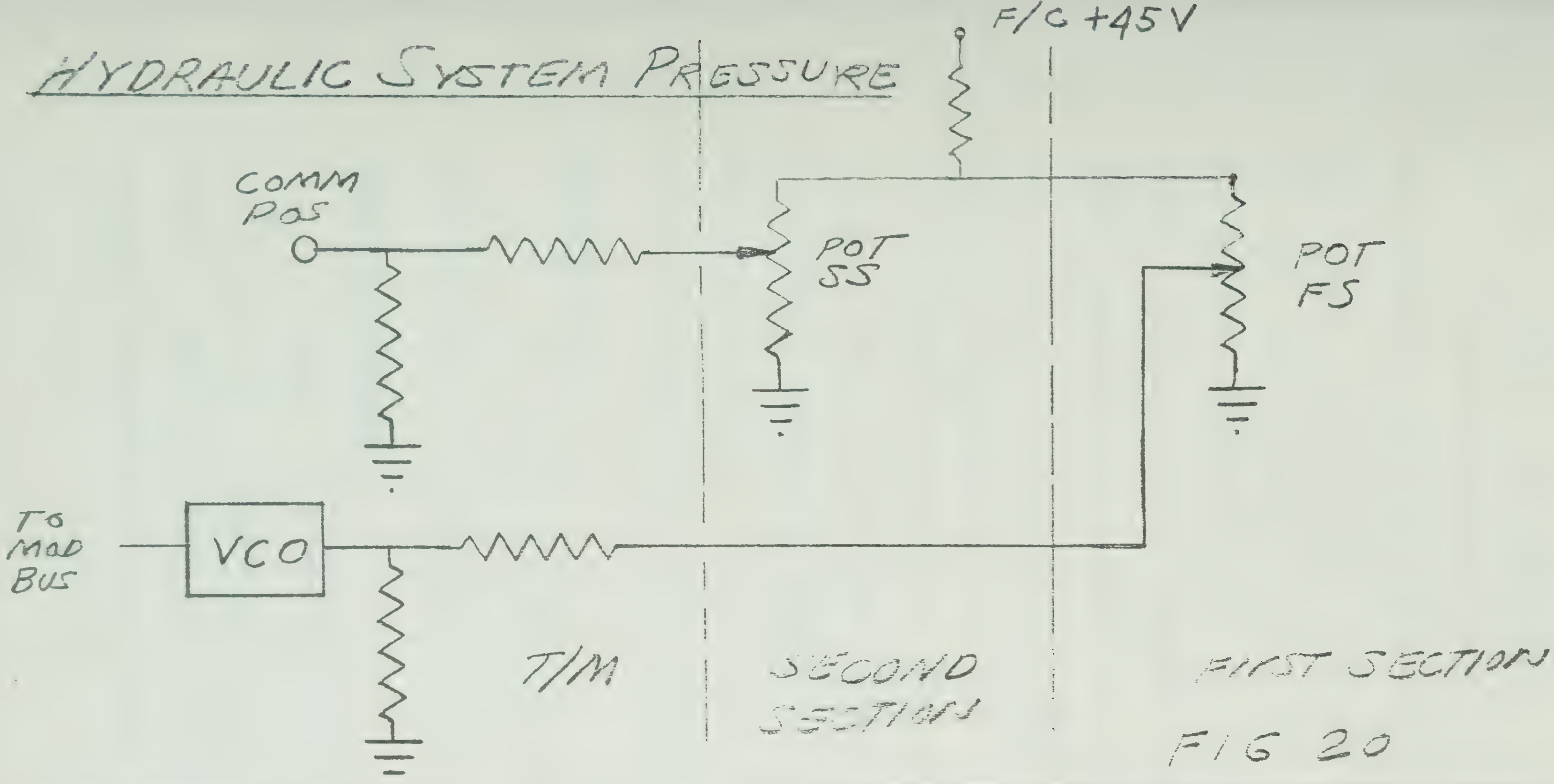
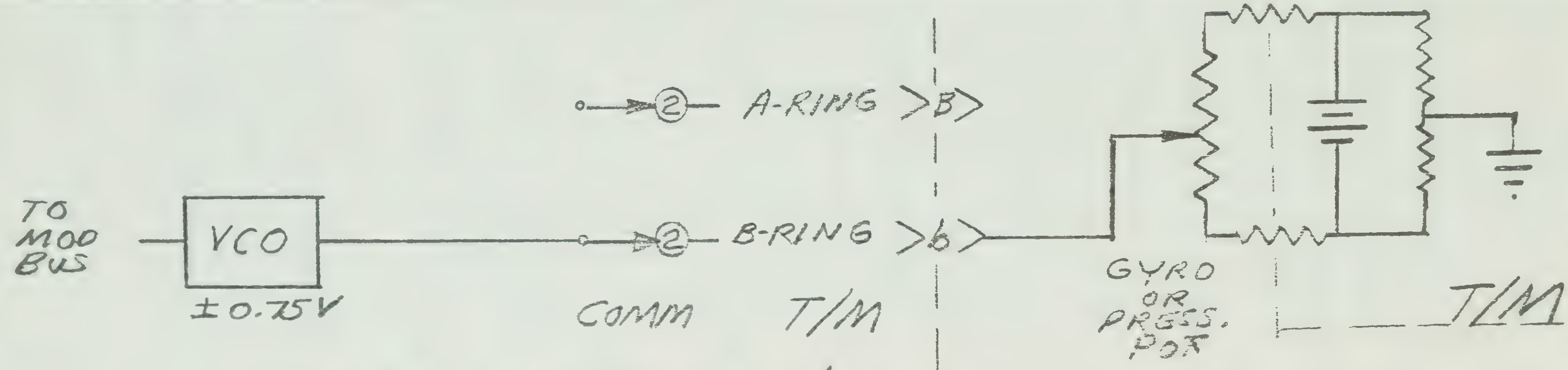


FIG 20



Potentiometer type transducer-Commutated
FIG 21

LMSD 450613
COPY NO._______

SC-3-b (6)

TOPIC: DUAL LOOP OSCILLATOR

DLO-1, -1A (Low Gain)
DLO-2, -2A (High Gain)

The transistorized Dual oscillator, shown in figure 1, was developed to work with a wide variety of transducers, supplying accurate measurement data under severe environmental conditions. The "dual loop" design makes the midband frequency independent of the transducer characteristics. The basic oscillator is the Wein Bridge type, noted for its excellent waveshape, and low zero-drift, important considerations in an oscillator used in a high-capacity multi-channel system, and operating under extremely low limits for intermodulation distortion. The Wein Bridge oscillator uses RC feedback and oscillates over a wide-band of frequencies simply by changing the RC time constant. The oscillator is used for Telemetry on channels 4 through 14, being limited on the low end mainly by the physical size of the condensers, and on the high end by the lead capacity to the transducers.

SPECIFICATIONS:

Table I lists the important electrical and mechanical characteristics of the DLO for both high and low gain version.

MODULATION ANALYSIS:

Figure 1 is a simplified block diagram of the Dual Loop Oscillator, drawn to emphasize the "dual loop" operation of the circuit. One loop consists of the RC network, which is the frequency determining portion of the oscillator.

The second loop is through a second bridge circuit made up, at least in part, by the transducer. When an external input such as pressure or strain is applied to the transducer, it causes a change in its resistance or reluctance and unbalances the bridge. The transducer makes up at least two arms of the bridge, increasing the unbalance signal (the input has opposite effects on the opposite arms). The unbalance signals or side-loop feedback will also vary in phase with the direction of bridge unbalance. Signal (EM) is fed to the summing point through a phase shift circuit which is adjusted so that EM is exactly 90°out of phase with E_2 at the summing point.

The vector sum, E, is thus shifted slightly, causing the oscillator frequency to change. The direction of the frequency shift depends upon the relative phase of Em.

THEORY OF OPERATION:

Figure 2 is a detailed block diagram of the Dual Loop oscillator. The output of the Wein Bridge oscillator is transformer coupled to the transducer. If a two arm transducer is used, as shown, the other two arms are precision wirewound resistors, mounted internally in the oscillator. The Cal Z (zero calibrate) switch is normally closed and couples the side-loop signal back to the Wein Bridge.

As the output signal is very small for small input signals (transducer bridge near null) the sensitivity must be increased by the addition of transistorized amplifiers. For variable reluctance transducers and potentiometers the single A-C amplifier (#2) is all that is required.

TABLE I

SPECIFICATIONS FOR THE DUAL LOOP OSCILLATOR:

	Reference Drawings
Schematic:	1033993, 1903473
Acceptance Test Specifications:	1700499, 1700500
Manufacturing Test Procedure:	
Set-up procedure:	

Mechanical

Size:	3.75 X 3.55 X .665 inches
Weight:	8 oz. (maximum)

Transistor Complement

	DLO-1, -2 DLO-1A, -2A
Q1	2N332
Q2, Q3, Q4, Q5, Q6	2N338
Q7, Q8	2N337

Electrical

Power Requirements	30 VDC 18 ma. (maximum)
Center Frequency	Channels 4 through 14
Frequency deviation	± 7.5% of Center frequency
Output voltage	Variable to 4.0V rms
Harmonic Distortion	Less than 2½% without filter. Less than 0.5% with filter.
Frequency drift	Frequency drift less than 0.3% of B.W per volt change in supply voltage Froquency drift less than ± 1% of B.W. at temperatures from 20°C to 85° C.
Linearity	Deviation from best straight line less than 1%
Sensitivity	±.1% change in r/R (one active arm) gives full B.W.
Output impedance	100k.
Freq. Change	3% of B.W.

Strain gage transducers, however, have a maximum output of approximately 14 millivolts and require additional amplification. To simplify manufacturing and reduce cost, all Dual Loop oscillators contain both transistorized amplifiers. The Low Gain (DLO-1) version is changed to the high gain DLO-2 simply by deleting jumper E and adding jumpers A, B, and F. The amplified signal is coupled through a phase shift circuit which produces the correct quadrature relationship for frequency (phase) modulation. The emitter-follower provides isolation and correct impedances.

The output of the frequency modulated Wein Bridge oscillator is fed to the Modulation Amplifier through a critically coupled, double-tuned band pass filter. This filter further reduces harmonics. A different filter is required for each channel.

The Amplitude Tell-tale, by shorting out part of the resistance divider in the output, lowers the amplitude of the output signal. This of course reduces the frequency deviation of the transmitter. At the event of interest, (such as first stage separation) the break-wire short is opened and the output rises to its normal value.

DETAILED CIRCUIT ANALYSIS:

The partial schematic of Figure 3 shows the Zero Calibrate Switch, one of the two dual A-C amplifiers, the quadrature circuit and the emitter-follower which couples the side-loop signal to the summing point in the Wein Bridge oscillator. The Cal Z switch (Q1) is normally "on", coupling the side-loop signal from the transducer bridge directly to the base of the first amplifier transistor.

The two lab set resistors, R2 and R3, are used to adjust the sensitivity or bandwidth of the oscillator. The amplitude of the signal coupled from the transducer, controls the frequency deviation of the oscillator. Maximum sensitivity is obtained by deleting R3 and shorting R2.

The A-C amplifier is a standard circuit and is feedback stabilized, the gain being determined primarily by the ratio of R25 to R24. The ratio of the two resistors is selected to give a two-stage gain of 100. As mentioned previously, the other amplifier(not shown) is required only with strain gage transducers. Its gain is stabilized at 20. Due to the use of feedback stabilization, no transistor selection is necessary for either amplifier. Besides the gain factor, the other difference between the two amplifiers is the output coupling; the first side-loop amplifier uses RC coupling in the place of T_1.

A standard phase-shift circuit couples the amplified side-loop signal to the emitter-follower. C11 is selected to give a 45° phase-shift with R29 for the particular channel frequency. A 90° phase-shift is required, but the other 45° shift occurs across the RC network at the summing point in the Wein Bridge oscillator.

The emitter-follower, Q6, performs two functions. It isolates the phase-shift circuit from the Wein Bridge oscillator and also provides proper operating impedance for both circuits to work into. The high input impedance of Q6 prevents loading (and phase-shift) of the quadrature circuit. Its low output impedance

(which makes R30 effectively only 50 ohms) accomplishes side-loop injection without inserting a large resistance in the bridge leg. A large resistance would make temperature compensation much more difficult.

Figure 4 is a simplified schematic of the Wein-Bridge oscillator, arranged to emphasize its bridge construction. Q7 and Q8 comprise a two-stage amplifier, amplifying the small signal appearing across the bridge balance points. Regenerative coupling between the output and the input is provided by R33, C12, R31-32 and C13. If R33C12 = (R32+R31) C13, then zero phase shift between output and input to the amplifiers occurs at a frequency of:

$$1/2\pi\sqrt{R31+R32)R33C12C13}$$

Amplitude of the oscillations are stabilized by negative feedback through CR5, which is a Zenner diode. Any increase in signal amplitude increases the current through the zenner and increases the negative feedback. The result is a constant amplitude of oscillations over a wide range of operating conditions. The circuit is also insensitive to power supply shifts. As the B+ shift appears across both legs, balance does not change and frequency holds constant. Excellent stability, waveform, and freedom from drift is thus obtained without resorting to selection of transistors.

Channel frequency is adjusted by selecting C12 and C13. The temperature characteristics of these condensers must be compensated for if the oscillator is to maintain frequency under extreme temperature variations. Compensation is provided by R31 and R32. The 10,000 ohms in the lower leg of the bridge is divided between a carbon film resistor (negative temperature coefficient) and

a wirewound resistor positive temperature coefficient.

One difference between the DLO-1, -2 and the newer DLO-1A, -2A, lies in the method of amplitude stabilization. The older version used a Thyride in the place of the Zenner diode. The Zenner diode, however, gives improved amplitude stabilization.

The output circuit of the Dual Loop oscillator is shown in the partial schematic of figure 5. Excitation for the transducer bridge is coupled through T2. A zero adjustment (R44) compensates for small inequalities in the transducer legs. R43 and R46, both precision 500 ohm resistors, makeup two arms of the transducer bridge. They are not required when 4 arm transducers are used. C19 is used for phase balance and is set across either arm as required.

In the output to the filter, R40 and R41 form a voltage divider with R39 and are lab set to give the correct amplitude for the output signal. The output of a Wein Bridge oscillator is constant over a wide range of frequencies, but the transmitter requires different amplitude modulation signals for different channel frequencies. In effect, the higher channels are "pre-emphasized" or allowed a slightly higher output - giving a slightly greater transmitter deviation. The output to the modulation amplifier goes through the bandpass filter which further improves the waveform and reduces the possibility of intermodulation effects between channels. The amplitude Tell-tale has been previously described.

IN-FLIGHT CALIBRATION OF THE DLO:

Basically, the in-flight calibration increases the confidence level by providing calibration checks just after the telemeter has been exposed to large vibration forces (such as after launch) or by calibrating at a time just before some critical measurement.

The circuit shown in the partial schematic of figure 6 provides a zero calibration or center frequency check on the DLO-1 and -2. During normal operation, the base of the top transistor (NPN) is positive and the switch is closed, coupling the side-loop signal to the amplifier. The lower transistor is a PNP and as its base is also positive it is off. When a zero calibration check is desired on the center frequency, an external 30 volt D-C pulse is applied to the Cal Z terminal. Both transistor bases are now biased negatively, the NPN switch opens the connection from the transducer and the PNP switch is "on", grounding the input to the side-loop amplifier.

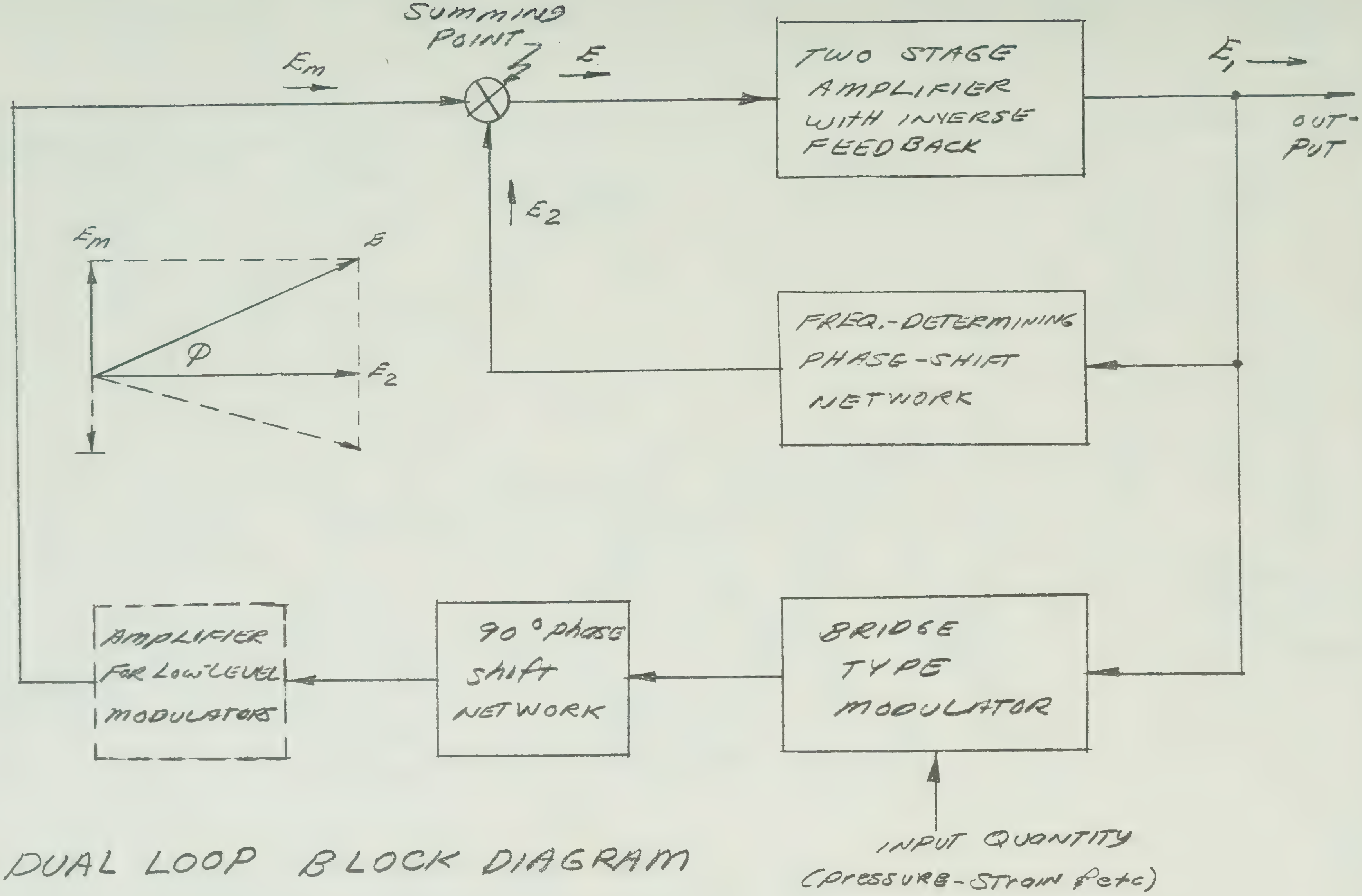
A center frequency check, however, will not detect changes in sensitivity or bandwidth, such as might result from a gross malfunction of some electronic component. Figure 7 shows how a plus and minus calibrate capability, checking both edges of the band, was added to the DLO-1A and -2A. The base of the Cal Z switch, Q1, is biased positive during normal operation. The switch is "on" and couples the side-loop signal to the amplifier.

During this time a small positive voltage is applied externally to zero Cal, + Cal, and -Cal points, back-biasing CR1, CR2, and CR3. CR4 is also back-biased from the +28 supply voltage, so all four diodes are "off".

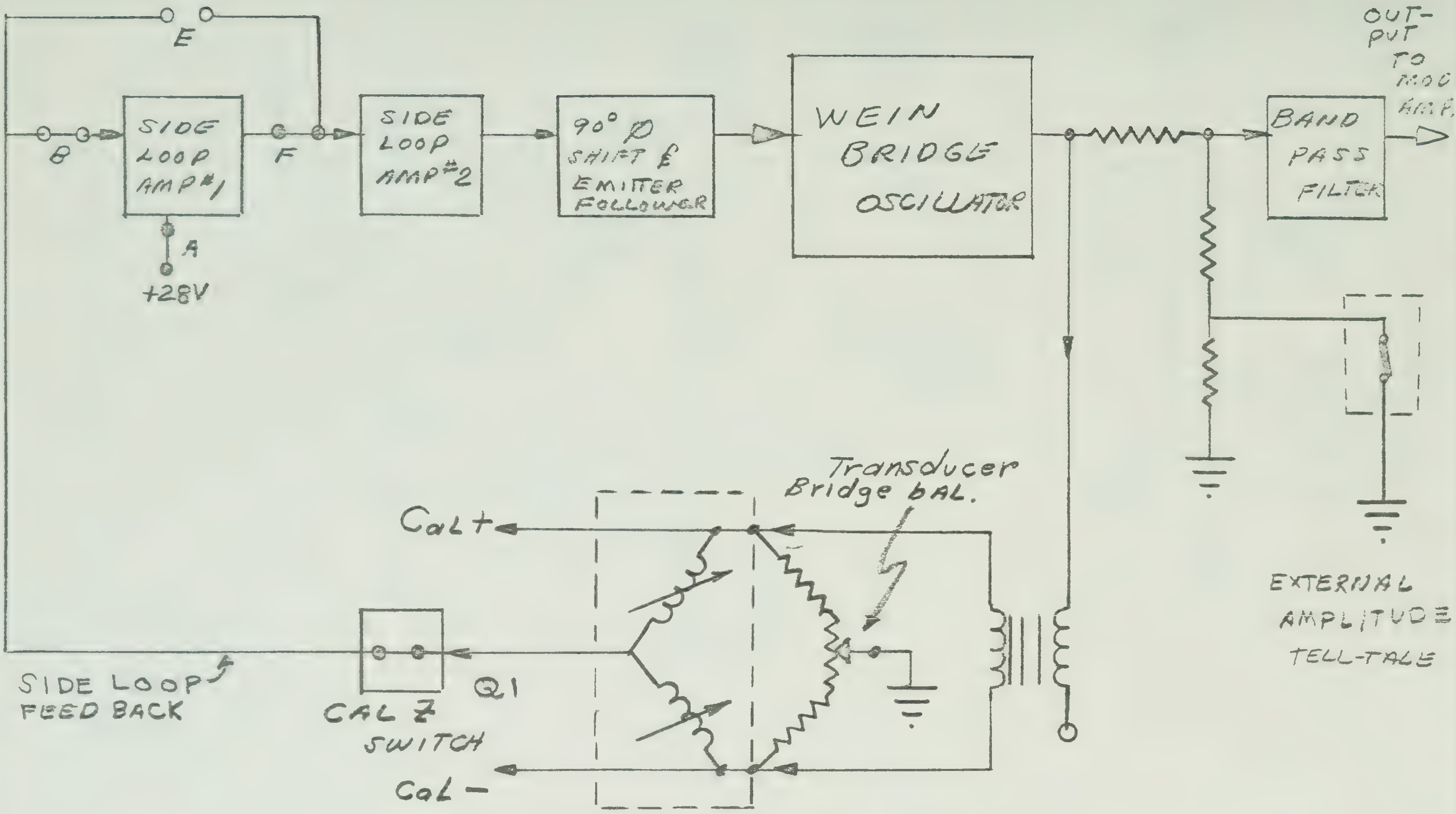
During the entire in-flight calibrate cycle an external -30 volts is applied to the Zero Cal terminal turning "on" CR1. A negative bias is thus applied to the base of Q1, opening the switch and grounding the input to the amplifier through a low value of resistance. The -30 volts Cal Z also turns on CR4, establishing a direct connection through it to the input of the amplifier. The negative voltage developed across R13, however, increases the back-bias on CR2 and CR3 and they remain off. As there is no input to the side-loop amplifier, the oscillator is now at its center frequency.

During the + Cal portion of the calibrate cycle, a -30 volts D-C is applied to the + Cal terminal, turning on CR2. One side of the transducer bridge excitation is now coupled directly through CR2 and CR4 to the amplifier. The voltage divider formed by R4, R9 and R13 adjusts the amplitude of the excitation signal to give maximum deviation of the oscillator frequency. As the phase of the signal is not exactly that which would normally be found across the balance points of the transducer bridge, C2 is added to correct the phase of the signal. The action of the - Cal portion is identical, except that CR3 is coupling the signal from the opposite side of the bridge, therefore the phase is reversed and the resultant oscillator deviation is in the opposite direction.

The actual cycling of the calibrate signals is, as shown in the figure, first plus, then zero, and then minus. The zero Cal is "on" whenever neither plus or minus are "on".


SUMMING POINT
E_m
E
E_1
OUT-PUT
TWO STAGE AMPLIFIER WITH INVERSE FEEDBACK
E_2
FREQ.-DETERMINING PHASE-SHIFT NETWORK
E_m
E
φ
E_2
BRIDGE TYPE MODULATOR
90° phase shift NETWORK
AMPLIFIER FOR LOW LEVEL MODULATORS
INPUT QUANTITY (pressure-Strain & etc)

DUAL LOOP BLOCK DIAGRAM

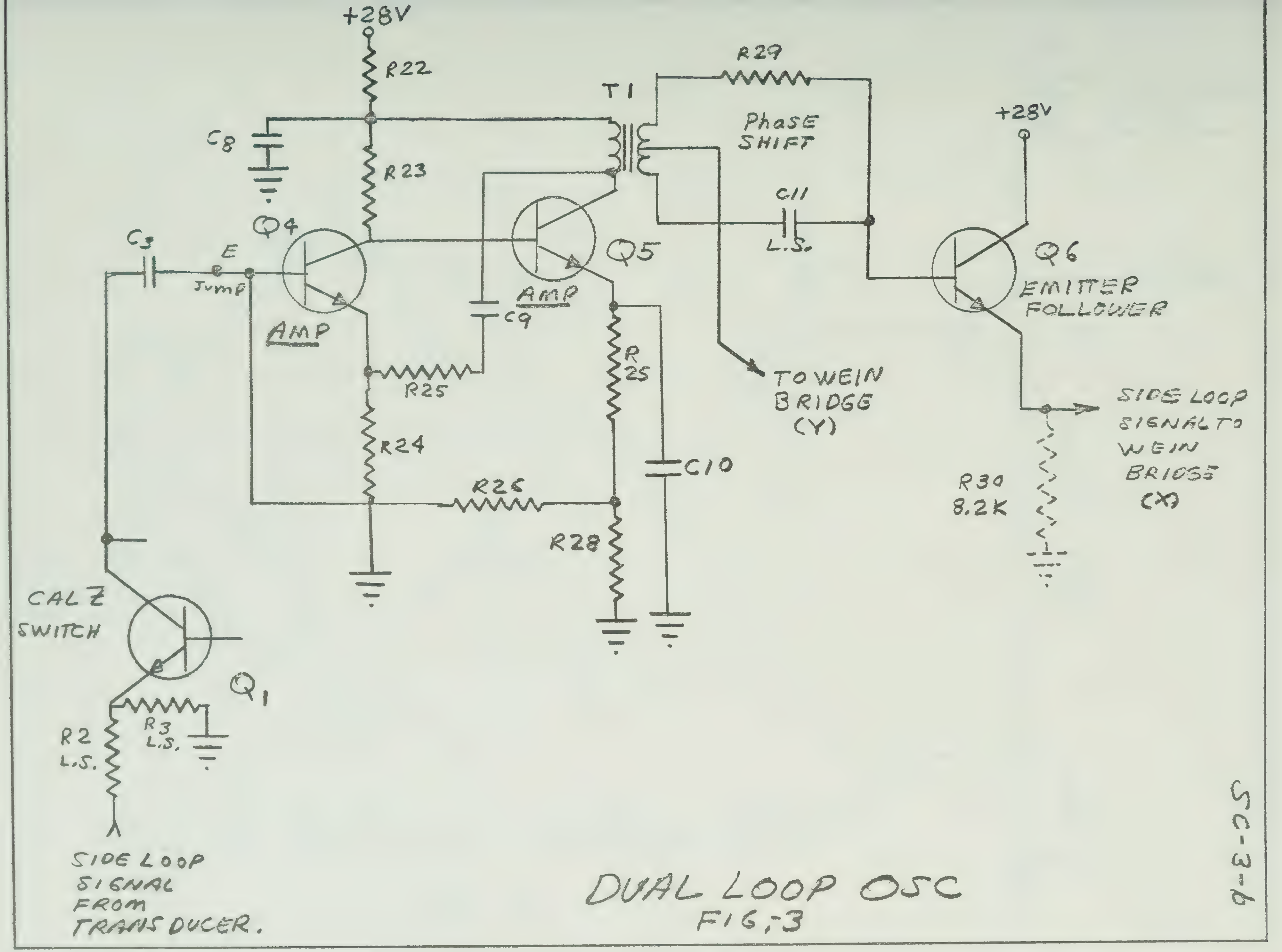


Detailed Block Diagram

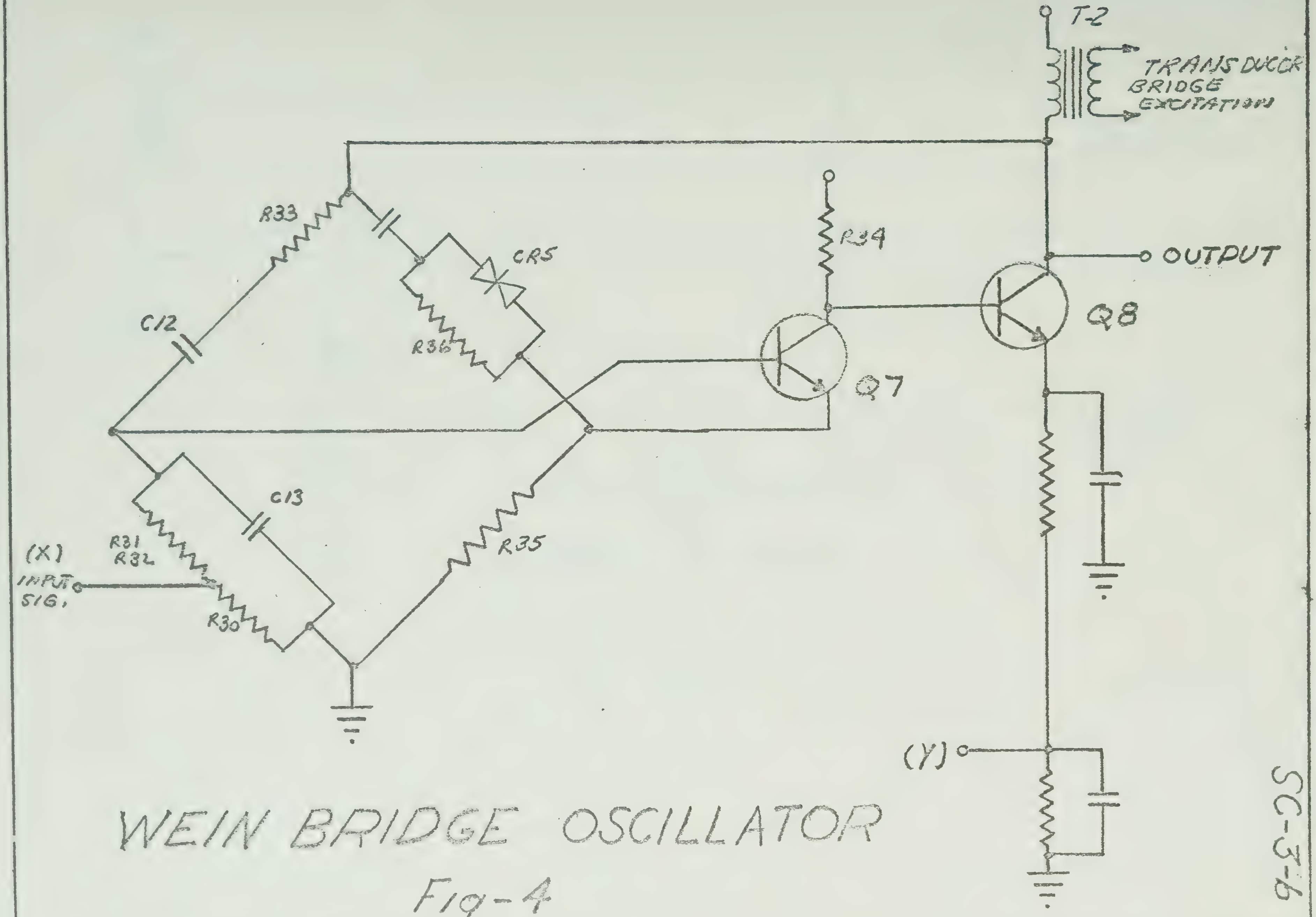
Fig-2

SC-3-b

SC-3-b



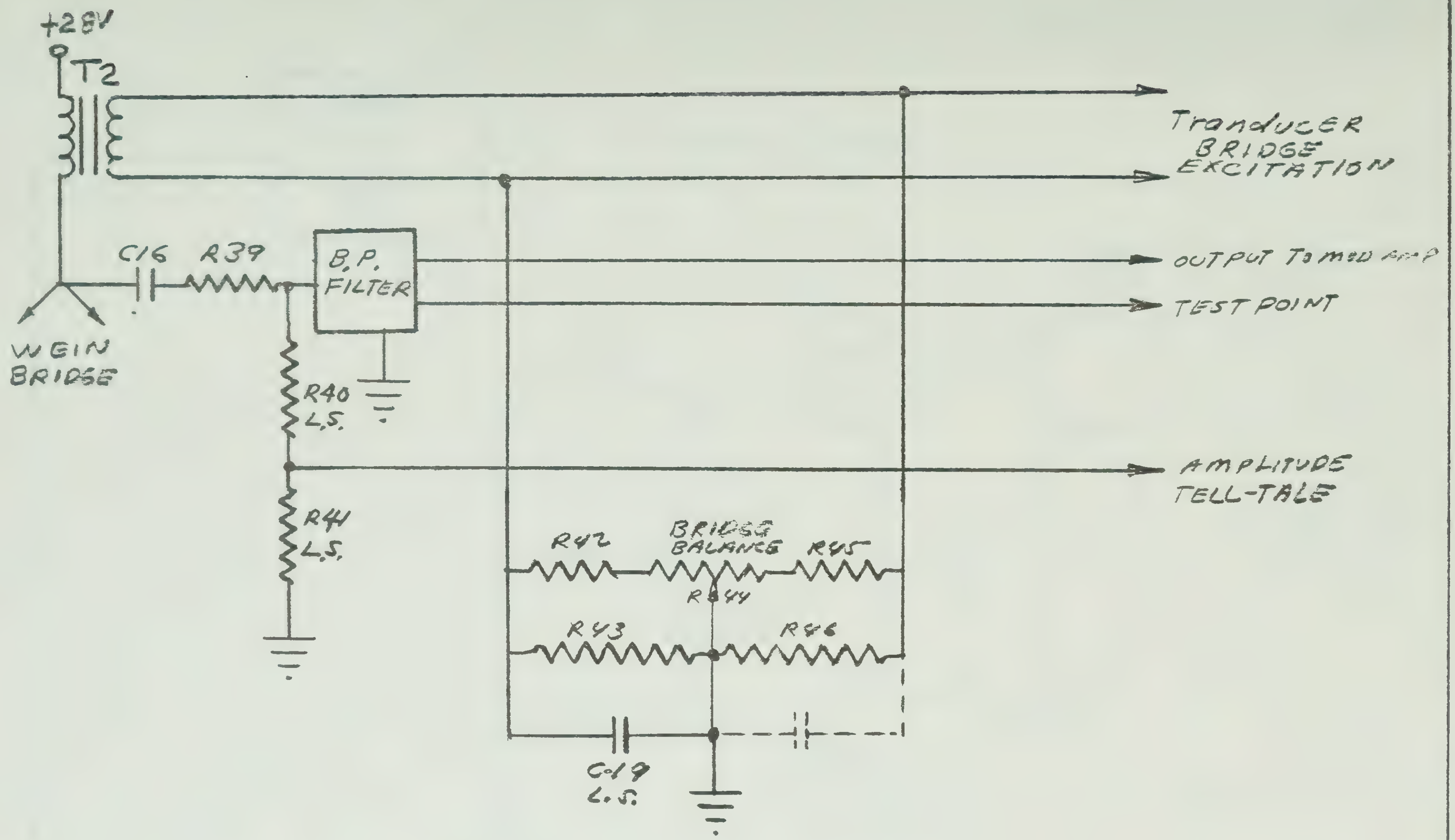
DUAL LOOP OSC
FIG.-3



WEIN BRIDGE OSCILLATOR

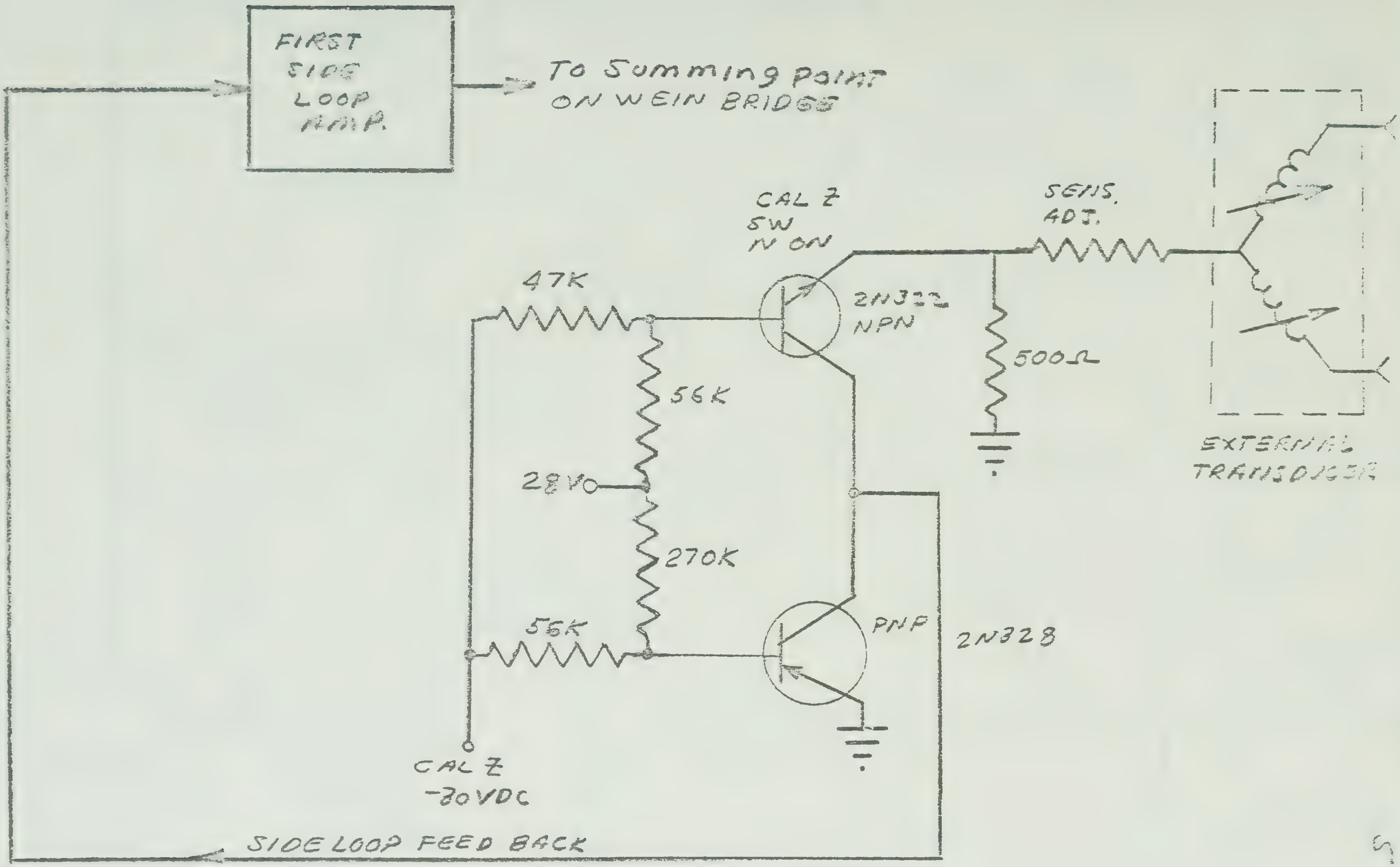
Fig-4

SC-3-b



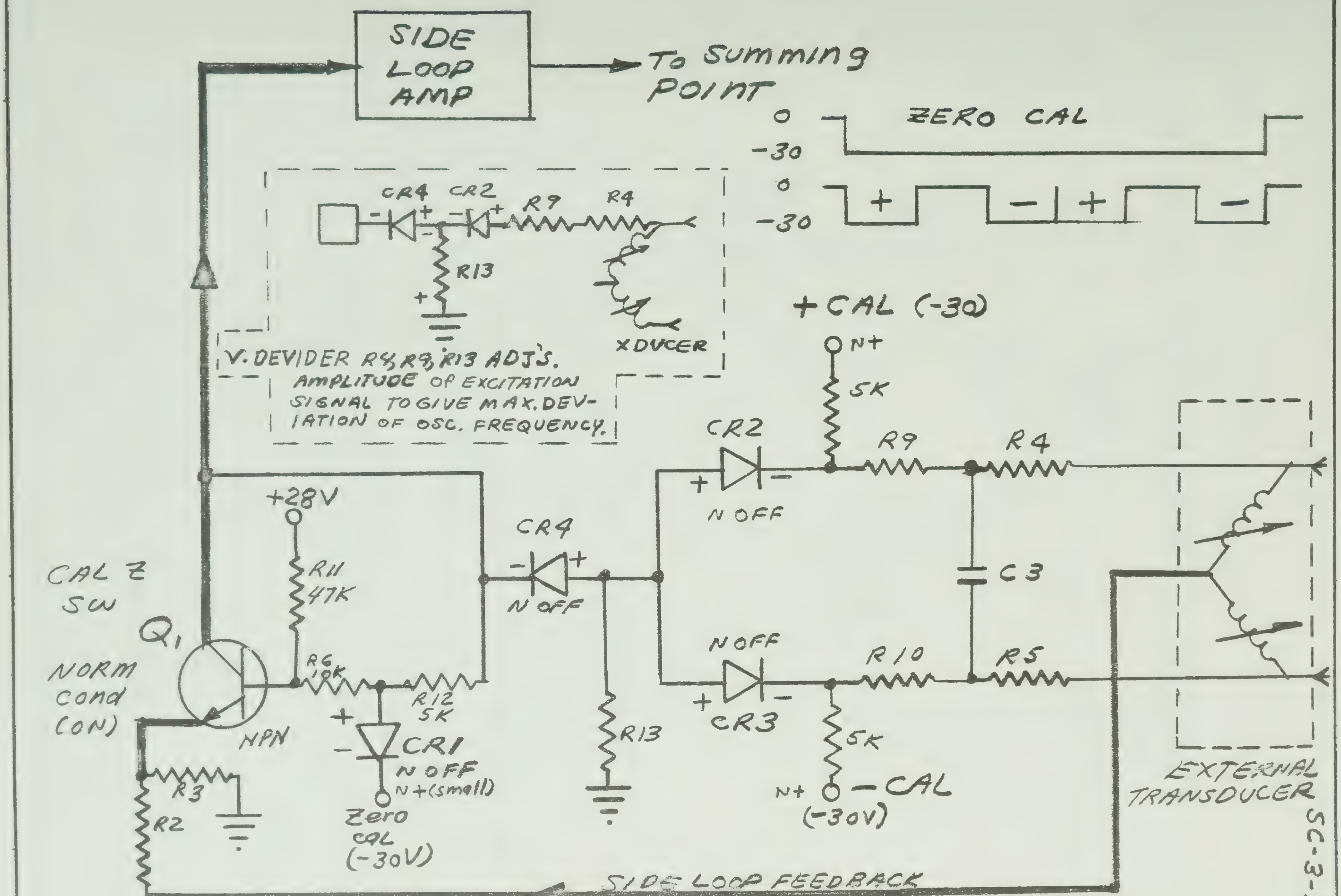
OUT PUT DLO

FIG-5



ZERO CAL. Circuit
FIG-6

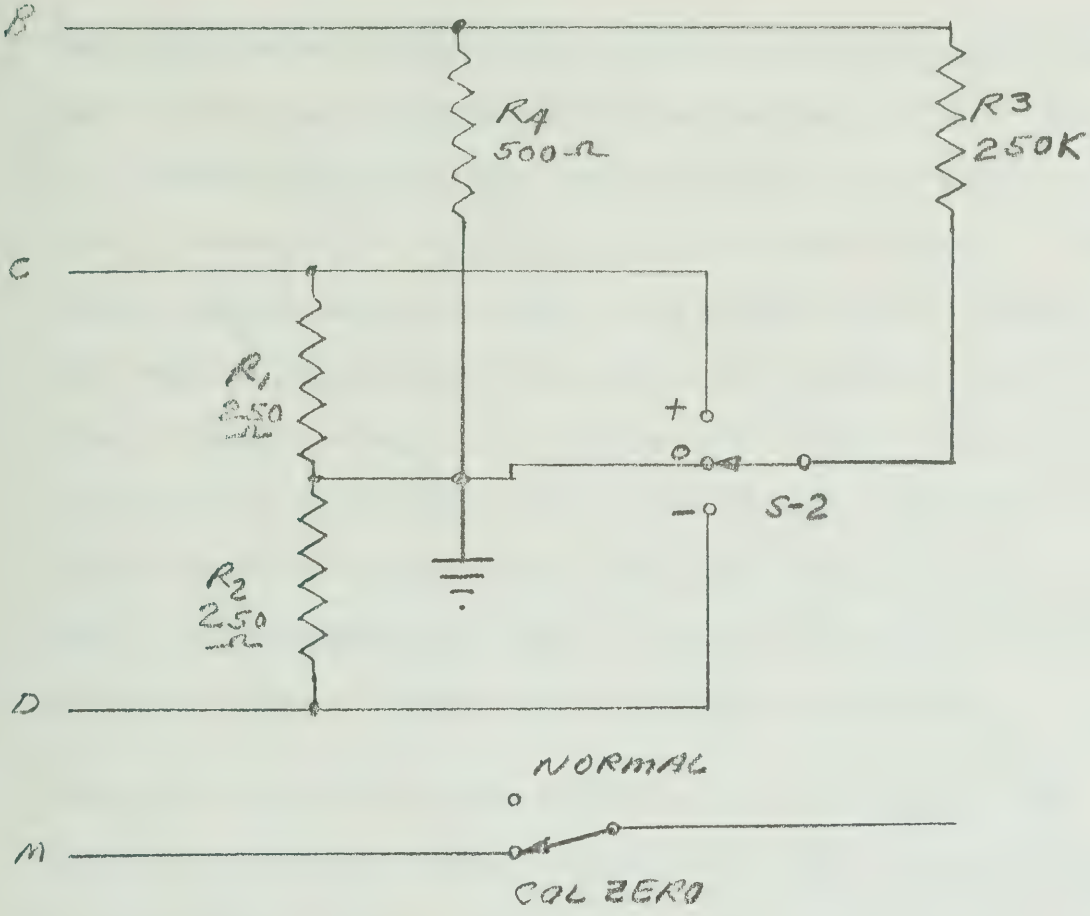
SC-3-b



V. DEVIDER R4, R9, R13 ADJ'S. AMPLITUDE OF EXCITATION SIGNAL TO GIVE MAX. DEVIATION OF OSC. FREQUENCY.

+, − & Z CAL CIRCUIT

SC-3-b



R1, R2	250 ohm deposited carbon bridge to be equal ±0.2 ohms
R3	250 K deposited carbon
R4	500 ohms deposited carbon

RESISTANCE BRIDGE SIMULATOR

DUAL LOOP OSCILIATOR
DLO-2

TOPIC: VOLTAGE CONTROL OSCILLATOR (VCO-3A)

The transistorized Voltage Control Oscillator was developed to combine ease of set-up and calibration with a maximum versatility. The VCO-3A is a blocking oscillator whose output frequency can be varied as a function of the amplitude and polarity of some control voltage. It thus operates equally well with either ac or dc input signals, giving it a wide range of applications. It can be used to directly monitor operating or supply voltages, or it can be used with any of a wide variety of transducers whose outputs are ac or dc signals proportional to the physical quantity being measured. The VCO-3A oscillates over a wide range of frequencies and is used on channels 3 through 18. It is used with high frequency pickups such as vibration transducers.

Set-up and calibration of the oscillator are exceptionally easy. The sensitivity of the oscillator is adjusted for 88% of bandwidth with standard voltage inputs. Any transducer can then be used with the VCO-3A, providing that the excitation voltage to the transducer has been adjusted so its maximum output is equal to the standard calibration input. As the calibration of the oscillator and the transducer are independent of each other, either component can be replaced in the field without requiring recalibration of the other.

The operating frequency of the oscillator is independent of transistor characteristics. Careful circuit design, along with temperature compensation of the RC networks, has achieved stable performance even in the critical POLARIS environment.

THEORY OF OPERATION

Figure 10 is a simplified block diagram of the Voltage Control Oscillator. The frequency control circuitry consists of a loop which is formed by the blocking oscillator, the transistor switch Q1, the RC network, the pick-off diode CR1, and the emitter-follower trigger-amplifier combination. In the normal course of its operation the blocking oscillator, which is free-running, sends a positive trigger pulse to the transistor switch Q1. This closes the switch and shorts condenser C. The voltage at the junction of R and C quickly falls to a negative value, cutting off CR1.

After the positive pulse from the blocking oscillator ends, the transistor switch Q1 cuts off again and condenser C starts charging toward a positive voltage. As soon as the voltage at the junction of R and C rises to a few millivolts above the input voltage, CR1 conducts and couples the rising voltage to the emitter-follower trigger-amplifier combination. Here it is amplified and inverted, appearing in the output as a large negative going signal. This signal is coupled to the sync input of the blocking oscillator. The free-running period of the blocking oscillator is 20% to 50% longer than the charge time of condenser C, so the arrival of the trigger pulse causes it to fire early. The triggering of the blocking oscillator initiates another pulse which goes to the transistor switch. This abruptly terminates the charging of condenser C, and thus starts another cycle.

The period between these cycles, is of course, a function both of the RC time constant of the frequency control network, and of the voltage at which CR1 starts to conduct. The output of the VCO-3A can thus be

frequency modulated by applying a control voltage (E_{in}) which will vary the bias on Crl. Figure 13 shows how the application of either a negative or positive signal to the input connection changes the period of oscillations. As the center-frequency of the VCO-3A is controlled by the RC network, shown in the simplified block diagram, and not by the free-running blocking oscillator, the frequency is thus independent of the characteristics of the transistors in the blocking oscillator circuit - hence no selection of transistors is required. The frequency will vary, of course, with supply voltage variations or with temperature. The two Zener diodes, CR3 and CR4, control the charging voltage (EB) and make it independent of supply voltage. Temperature compensation is used to eliminate RC drift with temperature.

The output of the blocking oscillator, as shown in the block diagram, is a sawtooth voltage which is coupled to a clipper-amplifier. The output of this clipper-amplifier is essentially a square wave, having very little even harmonic content. By eliminating the second harmonic, a more compact band-pass filter can be used.

DETAILED CIRCUIT ANALYSIS

Figure 11 is a partial schematic of the frequency control portion of the VCO-3A, including everything in the basic loop but the blocking oscillator. The two Zener diodes hold the voltage across the RC network to approximately 29 volts. When C4 is completely discharged, the voltage at the junction of R9 and C4 is -22v. C4 then charges toward a positive voltage whose maximum amplitude, (as set by R4), is 7v. C2 and C8 are filter condensers, preventing surges from either C4 or the blocking oscillator

from getting into the regulated voltage.

During the initial set-up of the oscillator, R4 is adjusted for 7v. The center frequency is then set by selecting C4, C20, R8 and R9 for the channel desired. After channel calibration is completed, R4 can tehn be used to adjust the frequency range for use with a certain type of transducer. If the output of a transducer, for example, is ±0.75v, then the R4 setting would not be changed. However, if the output of the transducer were 0 to +1.5v, then R4 would be adjusted so that center-frequency was obtained with an input signal of +0.75v. Figure 14 shows how R4 changes the frequency by changing the charging voltage.

R8 and R9 are a carbon film and a wirewound resistor, in that order, and are used for temperature compensation of the frequency control network. They also compensate for temperature effects in the two Zener diodes, both CR3 and CR4 having positive temperature coefficients. R7 provides a FINE adjustment on frequency, giving a maximum change of approximately 10% of bandwidth.

Since the base of Q1, the transistor switch, is at the same potential as its emitter, it is normally "off". When the 5v pulse arrives from the winding on transformer (T_1), Q1 is turned "on" and shorts C4, C20. Base current develops a negative bias across R11 and C5, which sharply cuts off Q1 upon removal of the positive 5v pulse. R10 is a current limiting resistor, added to protect the transistor.

The cathode of the pick-off diode, Cr1, is at the input signal potential. This diode is "off" until the voltage across C4 rises to a value slightly

positive in respect to the input signal potential. CR1 then conducts and couples the rising voltage, through C6, to the base of the emitter-follower Q2. The voltage across C4 continues to rise until it is between 20 MV and 40 MV positive in respect to CR1 cathode, at which time triggering of the blocking-oscillator occurs and C4 is discharged. The large negative spike following this small positive rise is due to differentiation action by the internal capacity of CR1. The discharge of C4 is coupled through this capacity (CR1 internal) to the base of the emitter follower shown on the waveshape as a sharp negative spike.

R13 provides a Sensitivity adjustment. During initial set-up a standard calibration voltage of ±0.75v is applied to the input and R13 is adjusted to give a frequency deviation of ± 44% of bandwidth. The Thermistor provides temperature compensation for sensitivity. Figure 15 represents a plot of output frequency versus temperature for three different calibration voltages. The change in frequency due to the temperature sensitivity of the circuit can be compensated for by raising the input voltage at higher temperatures. The thermistor, having a negative temperature coefficient, allows an increasing portion of the input signal to be applied to CR1 as the temperature rises.

Q2 is an emitter follower and provides isolation between the trigger amplifier and the frequency control network. It has unity gain and its output signal, developed across R16, is coupled through C7 to the base of Q3. Q3 has a high gain and injects a large negative going spike into the collector circuit of the blocking oscillator Q4. This injection of the triggering pulse is shown in detail in the partial

schematic of figure 12. The time at which triggering occurs is a function of the rise-time of the negative spike. Therefore it is desirable that Q3 have as high a gain factor as practical.

The trigger pulse is coupled, as a positive pulse, thru T_1 to the base of Q4, causing it to conduct. The voltage developed by current flow through R21, C10 and C11 in the emitter circuit, however, is large enough to cut-off Q4 at the end of the positive trigger pulse. The large pulse of current, which charges C10 and C11, biasing off Q4, also excites the tank circuit (T_1) into oscillation. The second half-cycle of oscillation is absorbed by CR2, and Q4 remains off. After cut-off, the voltage across C10 and C11 drops exponentially from its peak value until the next trigger arrives, when the condensers charge again. Thus a 3v peak to peak saw-tooth voltage is developed which is coupled through C13 to the squaring Amplifier Q5.

Q5 is biased at near saturation current and base current flows during the positive portion of the saw-tooth signal. Double clipping thus occurs, the base current cutting off the positive swing, and the negative swing cuts off Q5. The result is approximately a square-wave signal in the collector circuit, having an amplitude of 18v peak to peak. By squaring the wave-form, the even harmonic content is considerably reduced, enabling more efficient filtering with a smaller, more compact band-pass filter. [illegible] provides an output adjustment.

The Amplitude Tell-tale permits monitoring of some event (such as stage separation). The connection is normally shorted by a break-wire. By grounding one end of R31, the negative supply voltage is reduced by

the divider formed by R5 and R31. Upon opening of the break-wire, a greater supply voltage is applied across Q5, increasing the output signal 30% ±5%. This amplitude modulation of the output thus provides a record of event time.

SPECIFICATIONS FOR THE VOLTAGE CONTROL OSCILLATOR

Schematic: 1026070
Acceptance Test Specifications: 1700502
Manufacturing Test Procedure:
Set-up Procedure:

Mechanical

Size: 3.75 X 3.55 X .665
Weight:

Transistor Complement

Number	Type
Q1	2N696, 2N338, 2N332
Q2, Q3, Q4,	2N332
Q5	2N335

Electrical

Power Requirements: +28V dc, 10.0 ma. maximum
-28v dc, 4.0 ma. maximum (tell-tale grounded)

Center Frequency: Channels 3 through 18
Frequency deviation: ±7.5% of center frequency (Channels 3 - 18)
±15.0% of center frequency (Channels A - E)

Output Voltage: 100 MV minimum (channel 3 - 12)
(output loaded by 20K) 360 MV minimum (channel 13 - 18)
360 MV minimum (channel A - E)

Harmonic Distortion: 1.5% maximum

Frequency Drfit: Drift less than ±1.0% of B.W. for ±0.5v variation in either supply.
Drift less than ±3.0% of B.W. from 20°C to 85°C.

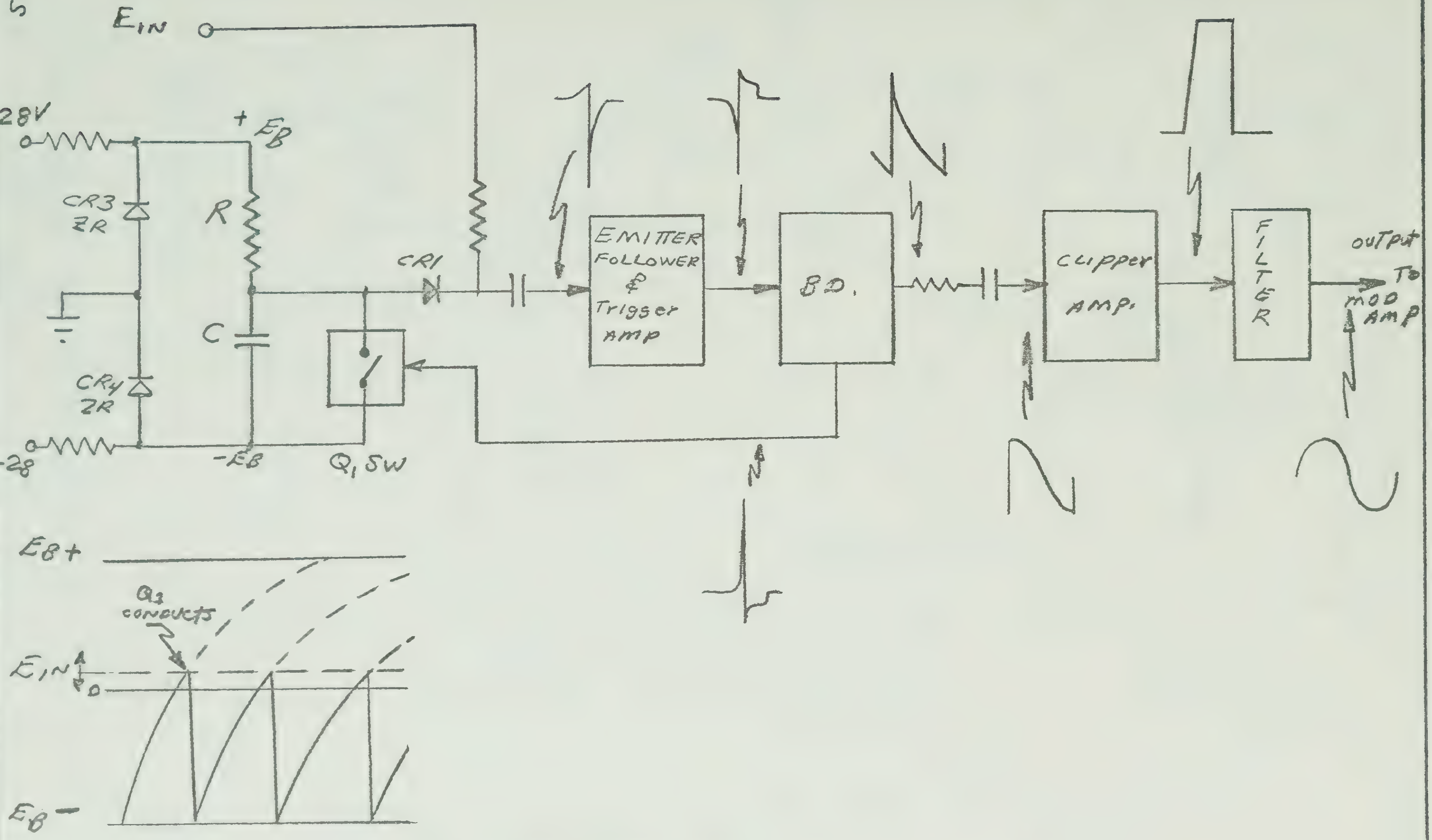
Linearity: Maximum of ±1.0% from best straight line.

Sensitivity: Output frequency shall deviate ±44 ±3% of bandwidth from center frequency when full range of input voltage is applied. (Normal operation is with ±0.75v, 0 to -1.5v, or 0 to +1.5v).

Input impedance: 200,000 ohms minimum
Output impedance: 100K nominal

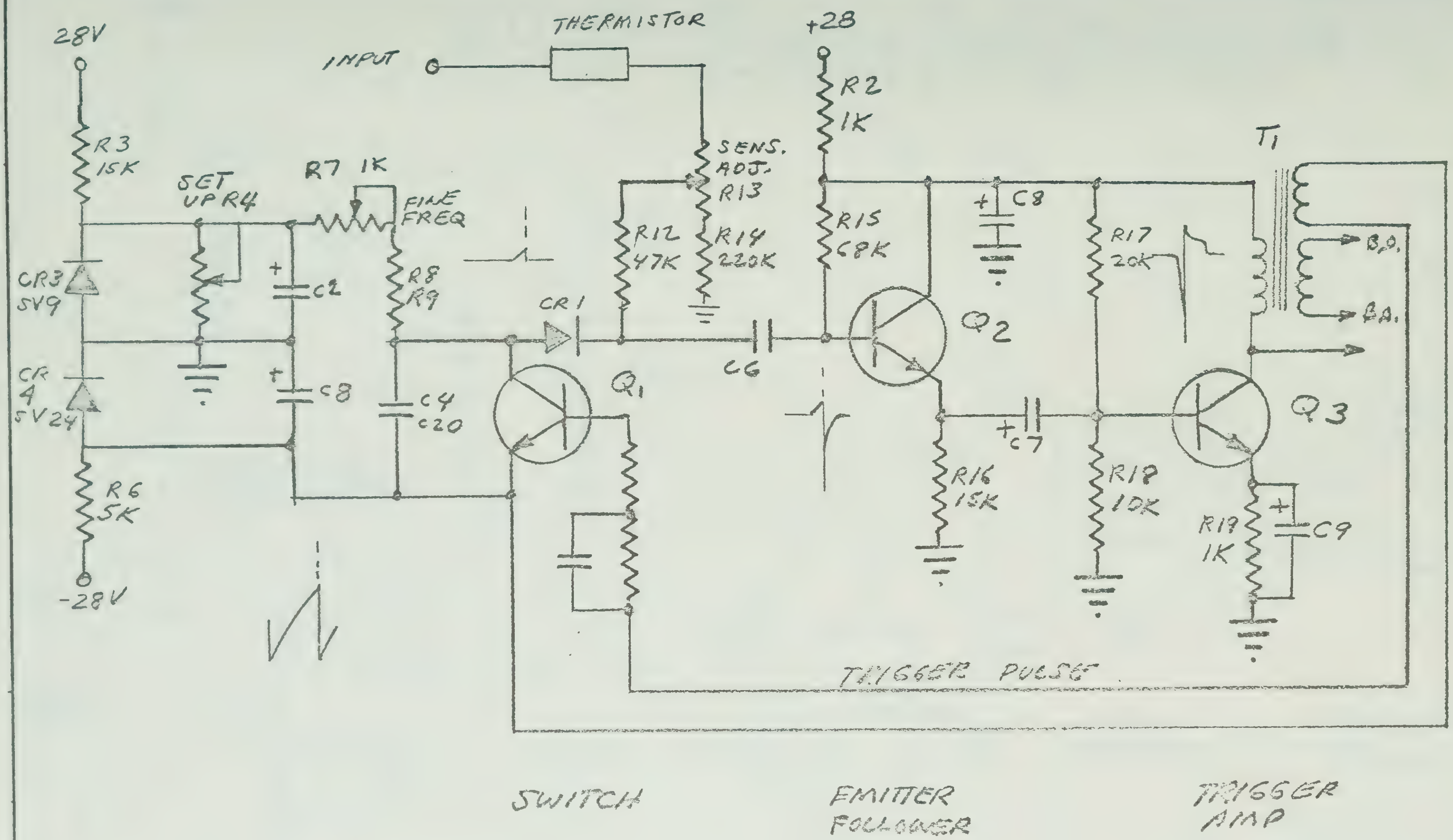
Amplitude Modulation: 16% maximum over band.

Tell-Tale Operation: Output decreases 30% ± 5% with Tell-Tale grounded.



Voltage Controlled Oscillator
FIG 10

SC-3-b



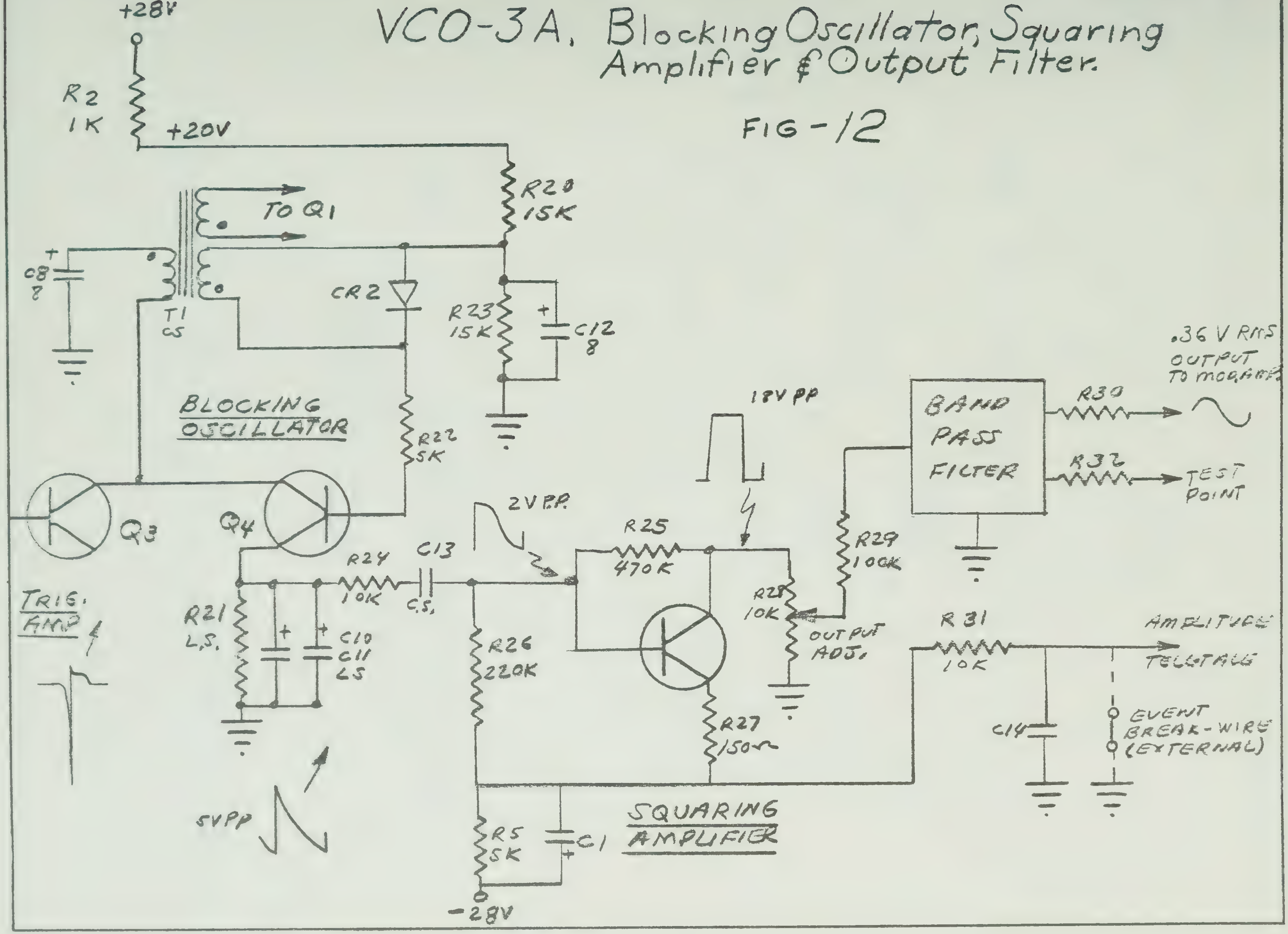
VCO — Frequency Control Network

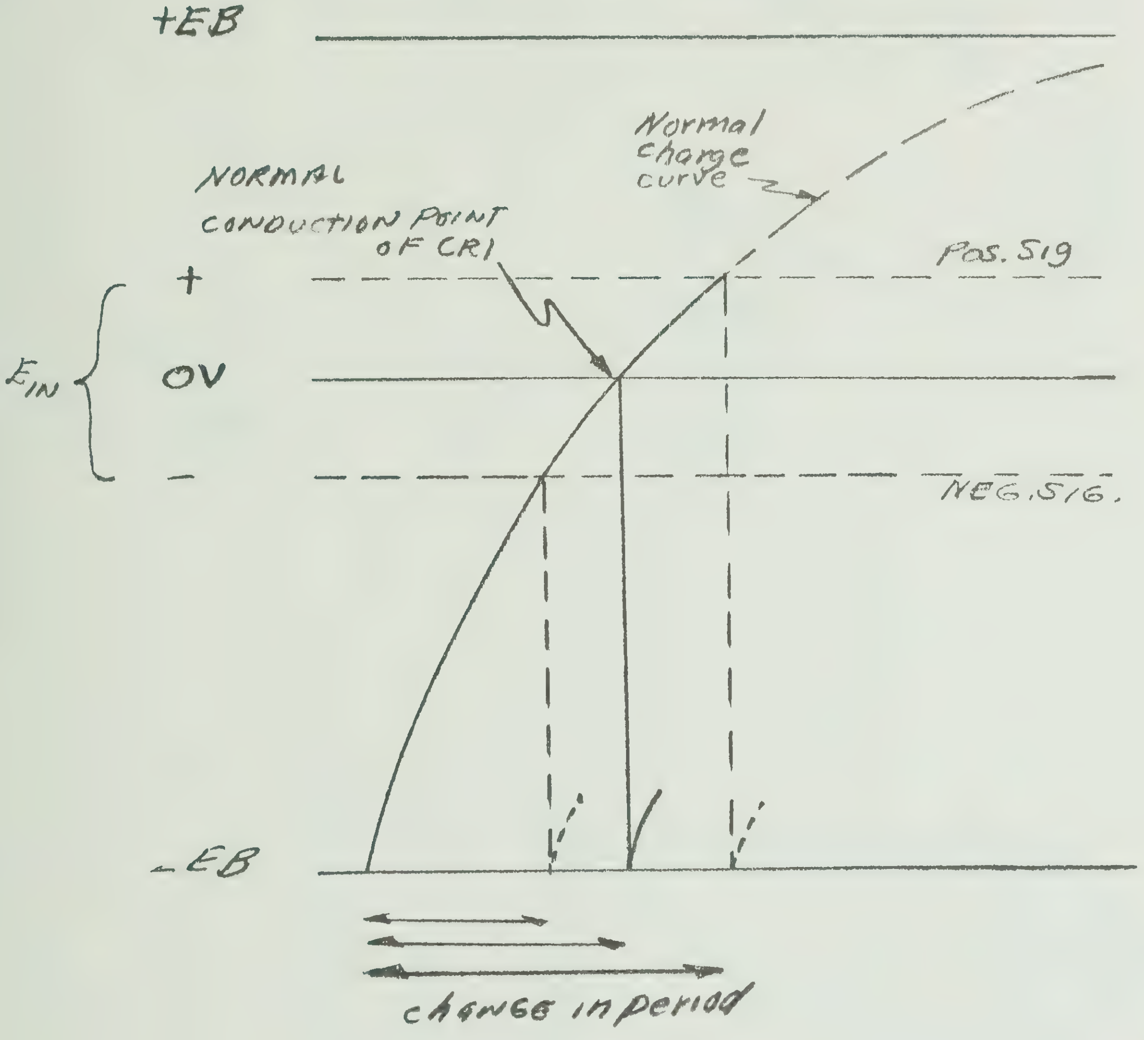
& Trigger Amplifier

FIG-11

VCO-3A, Blocking Oscillator, Squaring Amplifier & Output Filter.

FIG-12

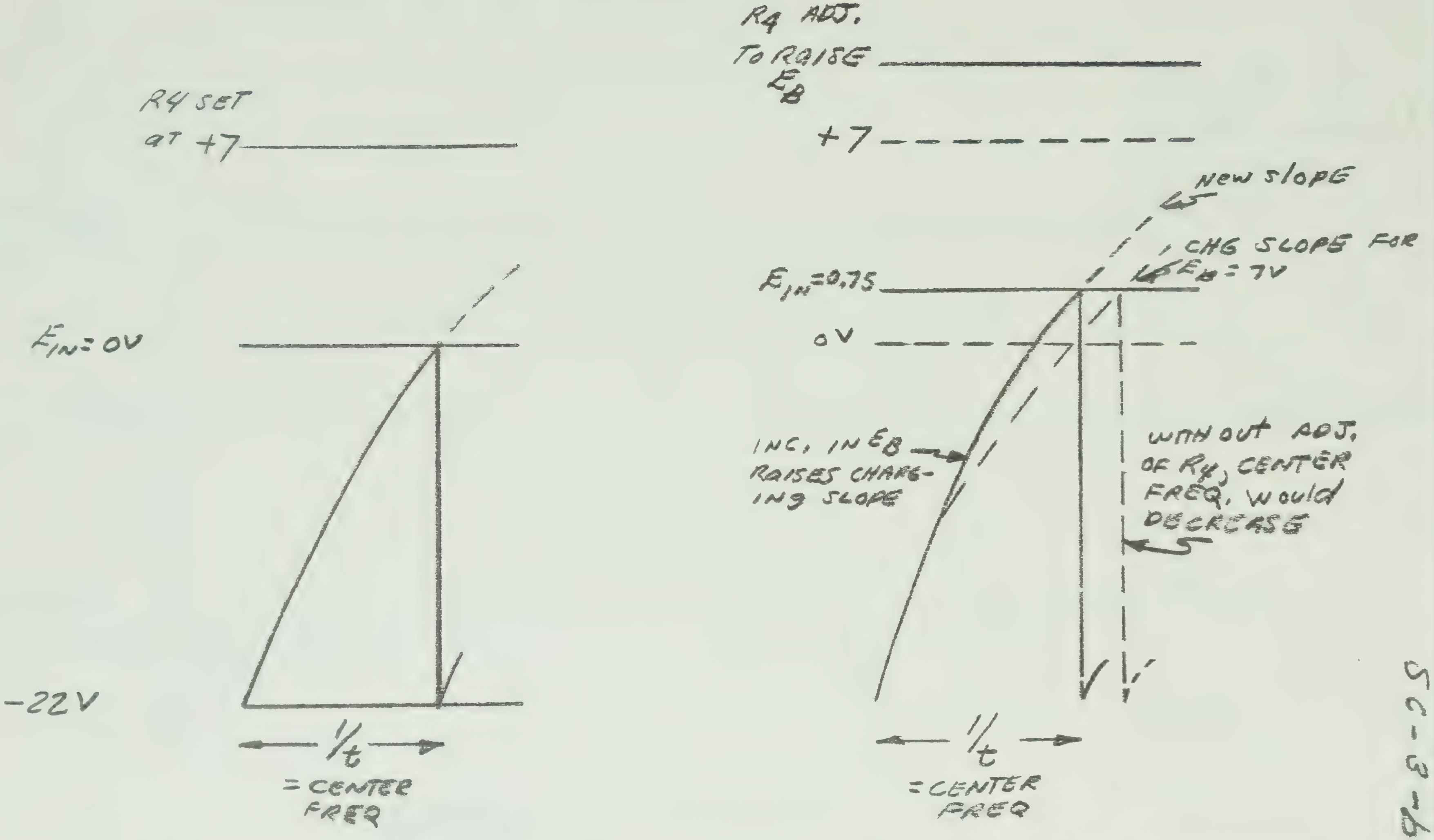




Frequency Modulation of VCO-3A

Fig-13

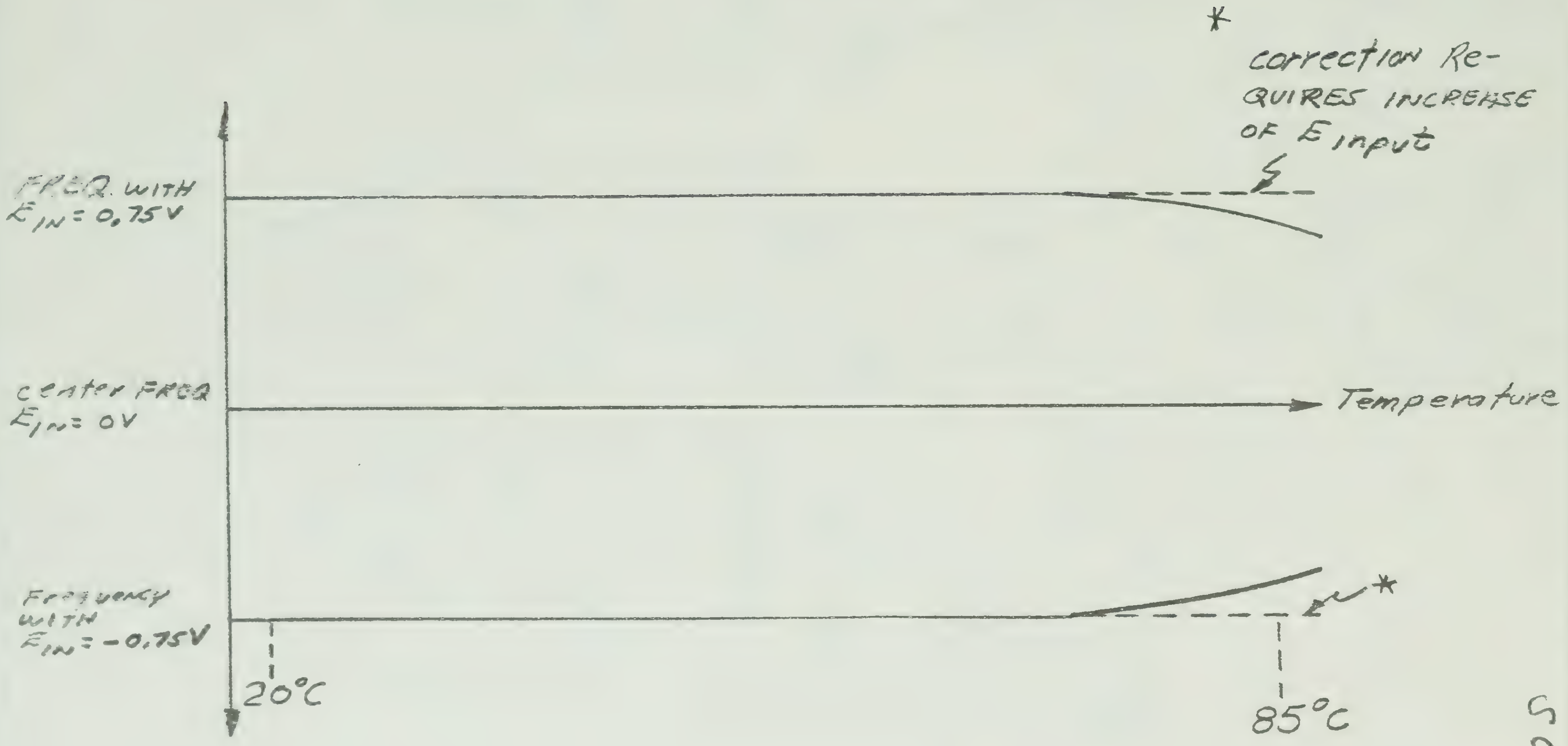
FIG.-1A



(A) SET UP FOR ±0.75 TRANSDUCER CENTER FREQ WITH E_{IN}=0V

(B) SET UP FOR 0 TO +1.5V TRANSDUCER CENTER FREQ WITH E_{IN}=+0.75V

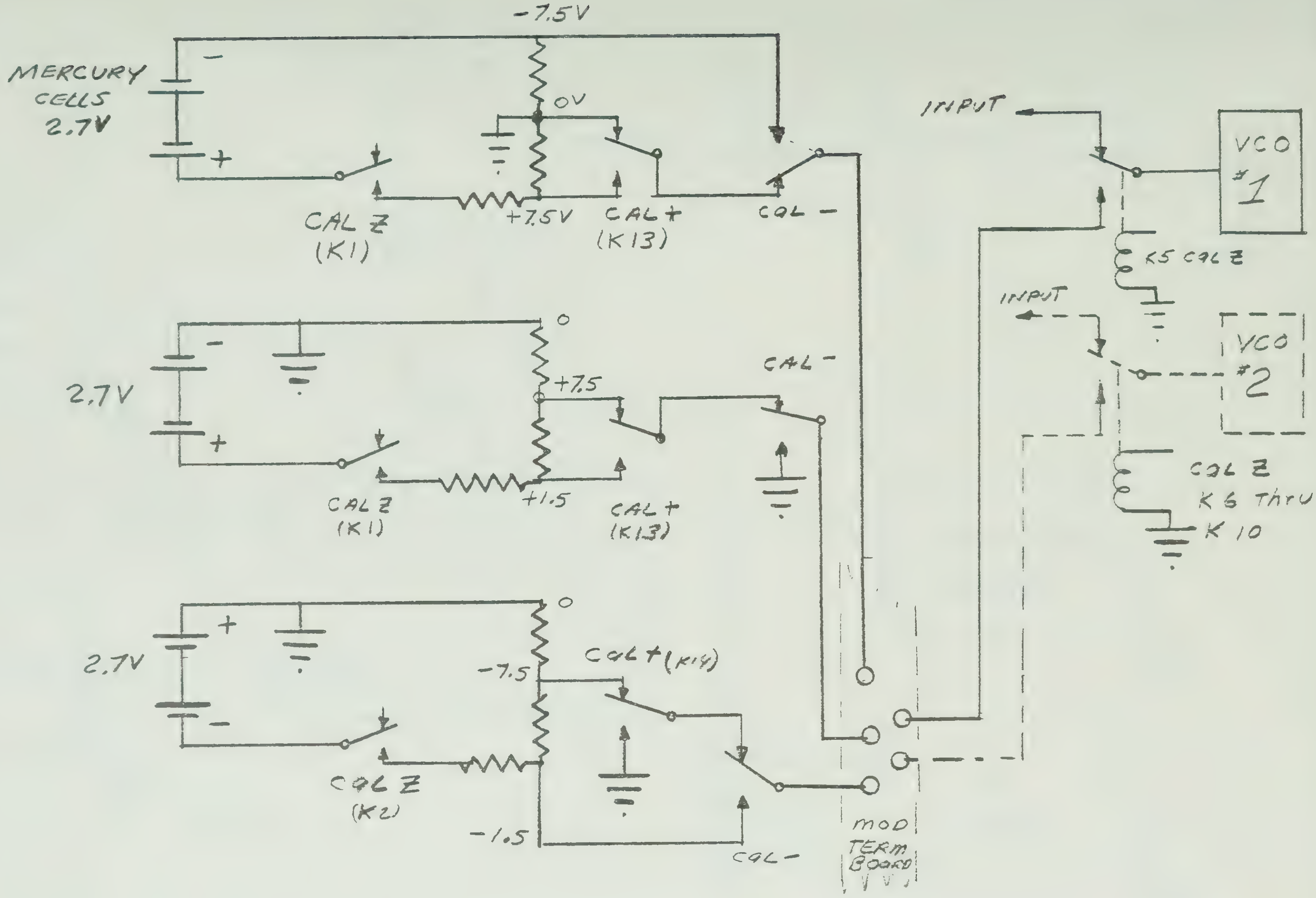
5C-3-b



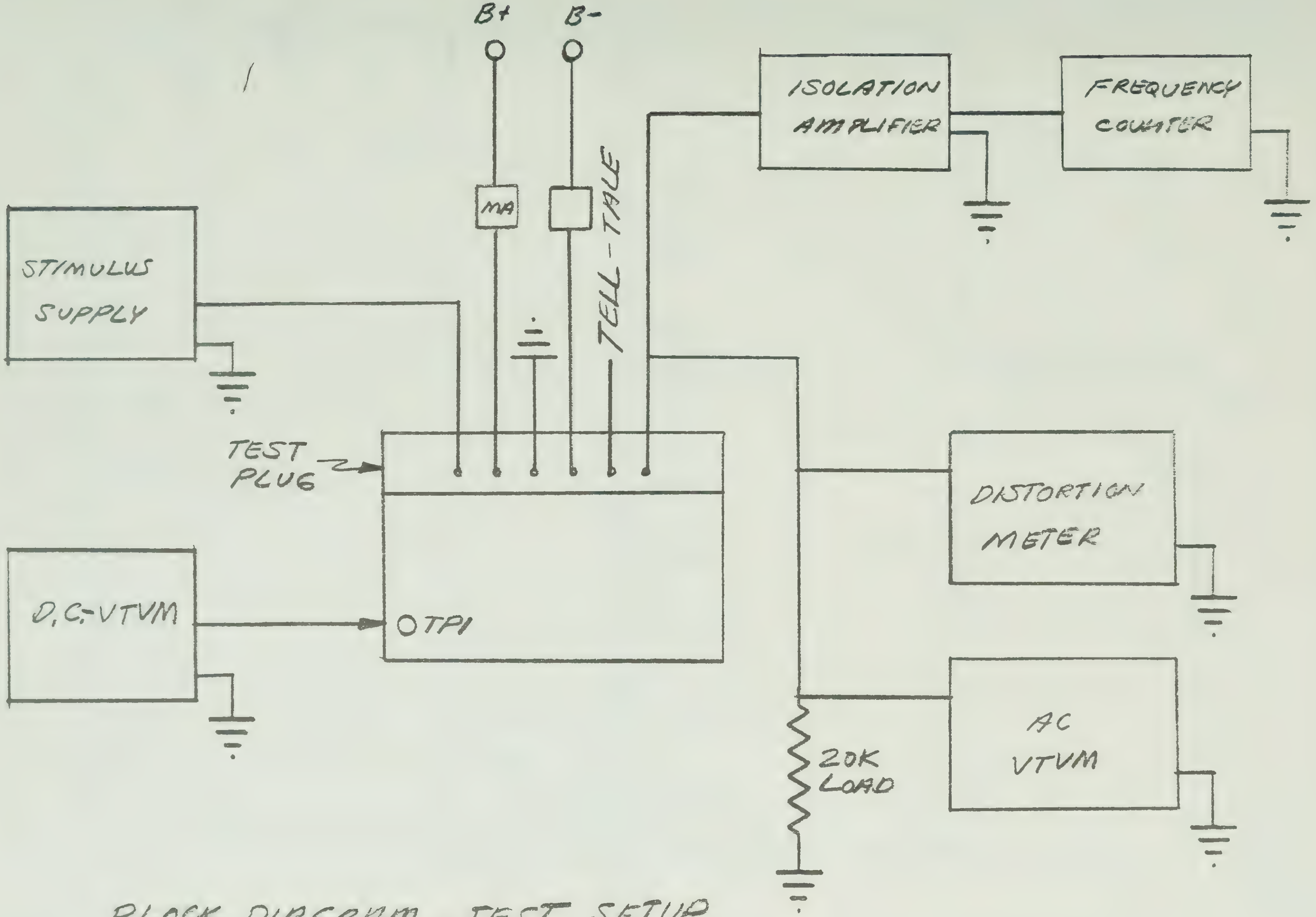
SC-3-b

CHANGE OF FREQUENCY WITH
TEMPERATURE.

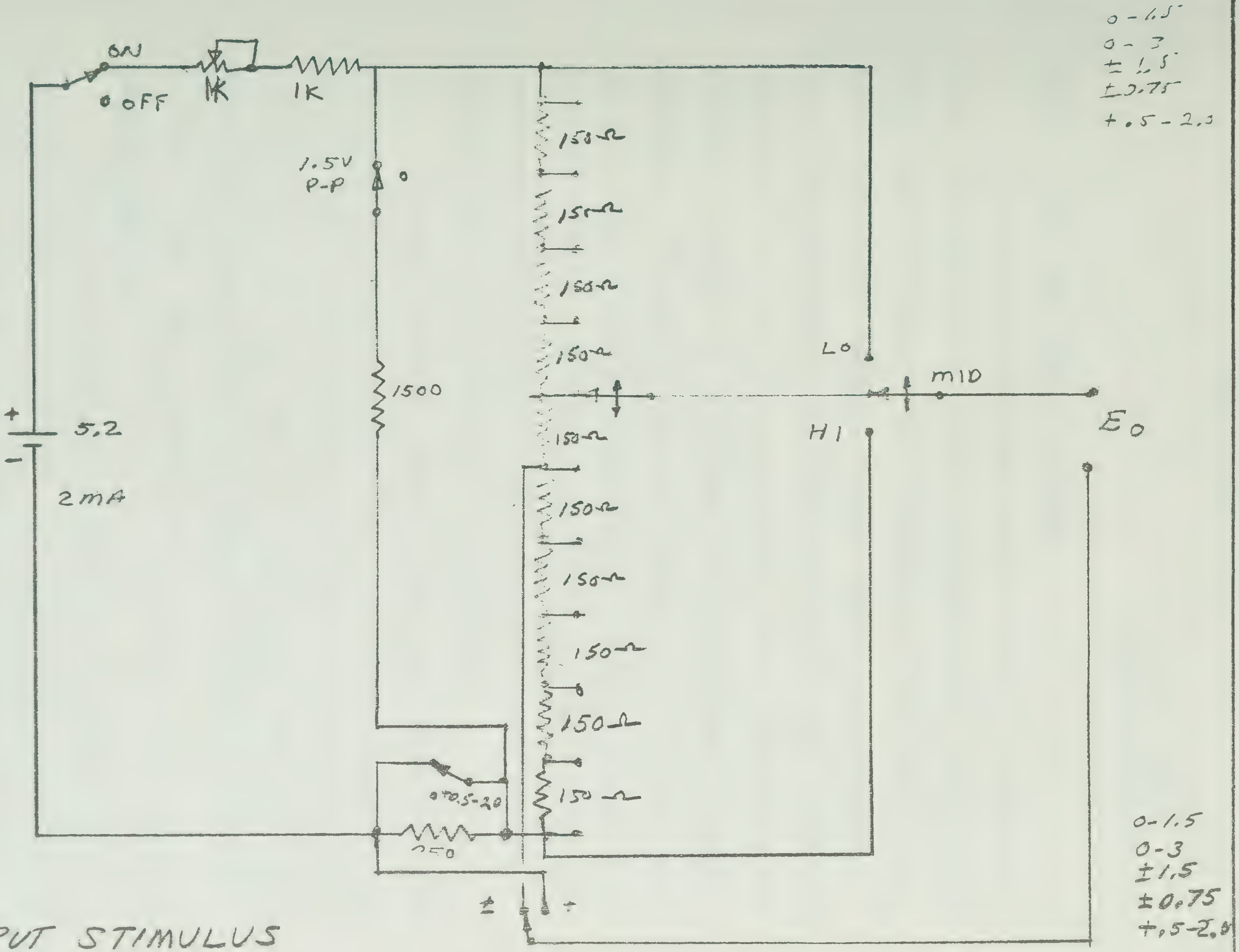
FIG-15



VCO CALIBRATE



BLOCK DIAGRAM - TEST SETUP
VOLTAGE CONTROLLED OSC.



INPUT STIMULUS
VC03A

LMSD 450613

LOCKHEED AIRCRAFT CORPORATION
MISSILE SYSTEMS DIVISION
XN NAVY TRAINING SCHOOL

Contract NOp-834
Course Notes SC-3-f

TOPIC: DUAL PHASE SENSITIVE CONVERTER

The transistorized Phase Sensitive Converter was designed to convert the output signals from gyro and accelerometers pickoffs to a proportional DC voltage suitable for modulating voltage control subcarrier oscillators. Examples of typical pickoffs are accelerometer or free gyro potentiometers and accelerometer or rate gyro differential transformers. The output of these pickoffs is an AC signal whose peak amplitude is proportional to the magnitude of some linear or angular displacement. The phase of this signal reverses with the direction of the input displacement. The Phase Sensitive Detector is essentially a synchronous detector. The output is a DC voltage whose amplitude is proportional to the amplitude of the 800 cps input signal, and whose polarity reverses with the phase of the input signal.

The Phase Sensitive Converter is small and compact, two units are mounted in a single standard module. The current requirements are low, both units drawing a maximum of 10 ma each. The output is stable with respect to time, temperature and supply voltage variations. The circuit also provides a high input impedance, effectively isolating the output circuit from the pickoff device.

THEORY OF OPERATION:

Figure 1 is a schematic of one of the two identical units in the Dual Phase Sensitive Converter. The input AC signal has a maximum amplitude of 1 VRMS, and can be any frequency from 400 cps to 2000 cps. It is coupled by C1 to the base of Q1. Q1 and Q2 are arranged in a

modified Darlington circuit and operate effectively as a single emitter follower. The net gain is slightly less than unity. The advantage of this arrangement is that it increases the input impedance of the transistor and prevents the circuit from loading the signal source. This is an important consideration as the gyro and accelerometer pick-offs are an integral part of the flight control servo loops.

The input impedance of the transistors is raised by increasing the effective alpha of the combined transistors. The base current (i_{b2}) of transistor Q2 is approximately related to the emitter current (i_{e2}) by the following expression:

$$i_{b2} = i_{e2} (1 - a_2) \qquad (1)$$

The higher the value of the alpha (a_2), the smaller the base current for a given output current. Examination of the figure shows that the emitter current (i_{e1}) for Q1 is supplied by the base current of Q2.

$$i_{b1} = i_{e2} (1 - a_1) \qquad (2)$$

Substituting equation (1) into equation (2) gives:

$$i_{b1} = i_{e2} (1 - a_2) \ (1 - a_1) \qquad (3)$$

Equation (3) can now be solved for the effective alpha of the two transistors:

$$a_t = a_1 + a_2 - a_1 a_2 \qquad (4)$$

If each transistor has an alpha of 0.95, the effective alpha of the two now becomes 0.9975.

The input impedance of Q1 can be estimated from the expression

$$Z_{in} = R3/(1 - a_t) \qquad (5)$$

where R3 is the 12K load resistor.

If the valve of alpha were only 0.95, the input impedance of the transistor would be 12K/0.05, or 240K. As this value is effectively in parallel with the 1.5 megohms of R1 + R2, the input impedance of the circuit is less than 240K. By raising the effective alpha to 0.9975, the input impedance of the transistor is now 12K/0.0025, or 4.8 megohms. The parallel combination with R1 + R2 now totals over 1 megohm.

The signal developed across load resistor R3 is coupled through variable attenuator R4 to the base of Q3. R4 is used to adjust the output of the Phase Sensitive Converter to ± 0.75 VDC when a 1 VRMS signal is applied to the input. Q3 provides the necessary amplification, raising the signal to the operating level required by the output circuit. As Q3 is the only transistor providing amplification, this stage determines the overall circuit stability. Temperature compensation and stable operation is obtained by careful design of the bias arrangement for Q3. First, I_c is considerably larger than delta I_{co}. Secondly, the ratio of R7 and R9 is adjusted to obtain maximum stage stability. If, for example, a supply voltage fluctuation tends to increase the collector current (i_c), a portion of the resultant voltage drop at the junction of R6 and R7 is coupled to the base of Q3. The drop in bias acts to oppose the change that caused it, and by proper selection of R7 and R9, a high order of stability can be obtained. It is necessary, however, that Q3 have a beta greater than 60. Final temperature compensation for the entire circuit is obtained by selection of R5 and R6, a carbon film and a wire wound resistor, respectively.

C4 and R7 provide a phase adjustment, permitting the amplified signal to be corrected for small phase shifts occurring within the stage and T1. This adjustment will change, of course, with different input frequencies.

The amplified signal from Q3 is coupled through the center-tapped transformer, T1, to the synchronous detector in the output circuit. Q4 and Q5 are PNP transistors and are operated as switches. They are turned on by application of a negative signal to their base. Current will flow in either direction through the transistor once it has been turned on. The switching voltage comes from the reference signal which is coupled through T2. As opposite ends of the transformer, T2, have opposite polarities relative to the center-tap, only one of the two transistors will be on at a particular instant. The reference signal originates from the same source that supplies the input signal from the gyro pickoffs. The reference signal thus has the same frequency and is either in phase or out of phase with the input signal. This phase relationship determines the direction of current flow through the transistors and the load resistor, R13.

Figure 2 is a simplified representation of the synchronous detector. The two transistors are represented as switches which are closed whenever the reference signal applied to that base is negative. Figure 2 (A) shows the current paths for both half cycles of the input signal when the input and reference signals are in phase. The polarity of the output DC voltage is positive. Figure 2 (B) shows the current paths when reference and input signals are out of phase. The polarity of the DC output is now negative. The output DC is smoothed by a Pi type filter, as shown in the schematic. The total ripple is less than 7.5 millivolts, peak to peak.

LMSD 450613
SC-3-f (5)

SPECIFICATIONS FOR THE PHASE SENSITIVE CONVERTER

Schematic: 1029774
Acceptance Test Specifications: 1700501

MECHANICAL

Size: 3.75 x 3.55 x 0.665 inches
Weight: 5.3 ounces

TRANSISTOR COMPLEMENT

Q1, Q2	2N328
Q3	2N335
Q4, Q5	2N496

ELECTRICAL

Power Requirements: 28 VDC, 10 ma. per unit. Reference signal input is 800 cps, 115 VRMS, 0.5 ma at primary.

Input: 800 cps, 1 VRMS, either phase.

Output: ± 0.75 VDC, polarity reverses with phase of input signal.

Maximum unbalance, 10 millivolts.

Stability: Temperature: Change in DC output voltage not more than ± 2% from 10° C to 85° C.

B+. Change in DC output not more than ± 1% for a ± 10% change in the 28 volt supply.

Time. Change in DC output less than ± 1.5% in 60 min.

Linearity: Input AC voltage (either phase) vs. DC output voltage is within ± 3% of the straight line connecting the end points.

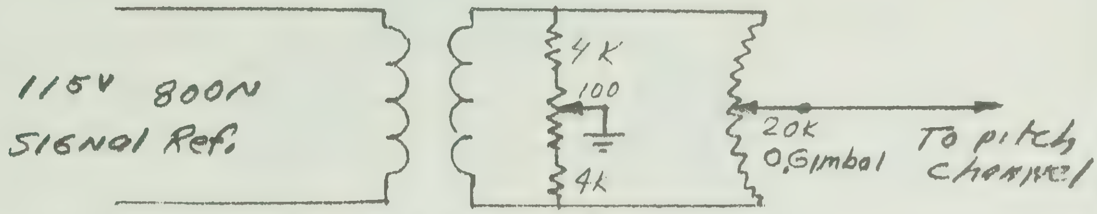
Input impedance: Minimum 1 megohm

Frequency Sensitivity: DC output will not change more than ± 0.5% for ± 2.5% change in input frequency.

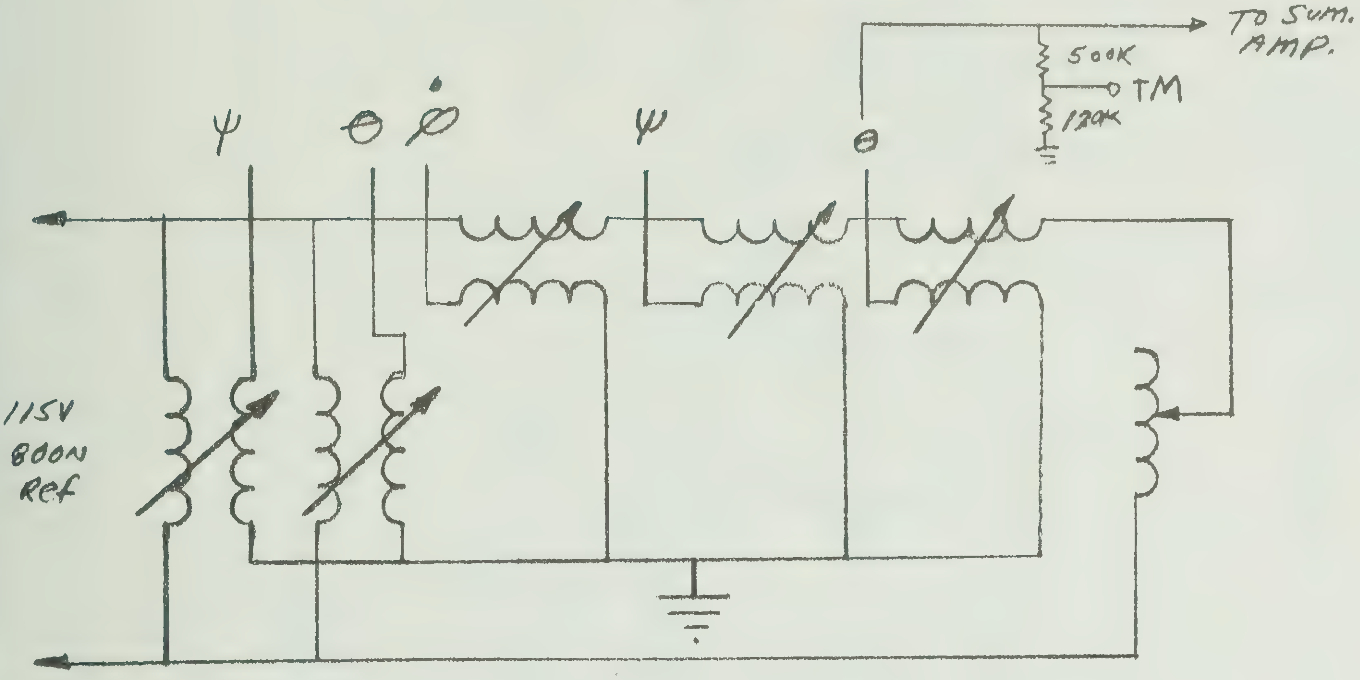
Rise Time: Nominal 10 milliseconds

Ripple: Output ripple 7.5 MV (peak to peak) maximum.

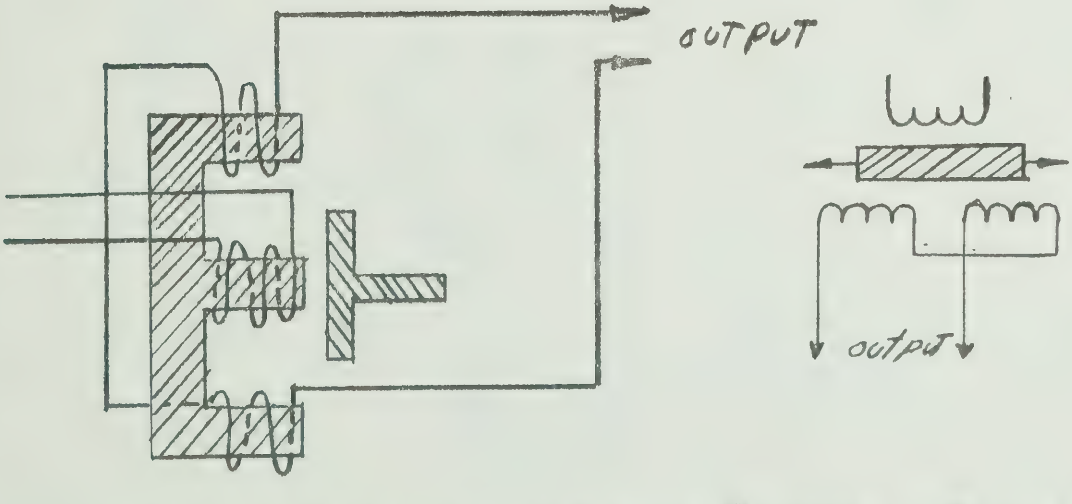
SC-3-F



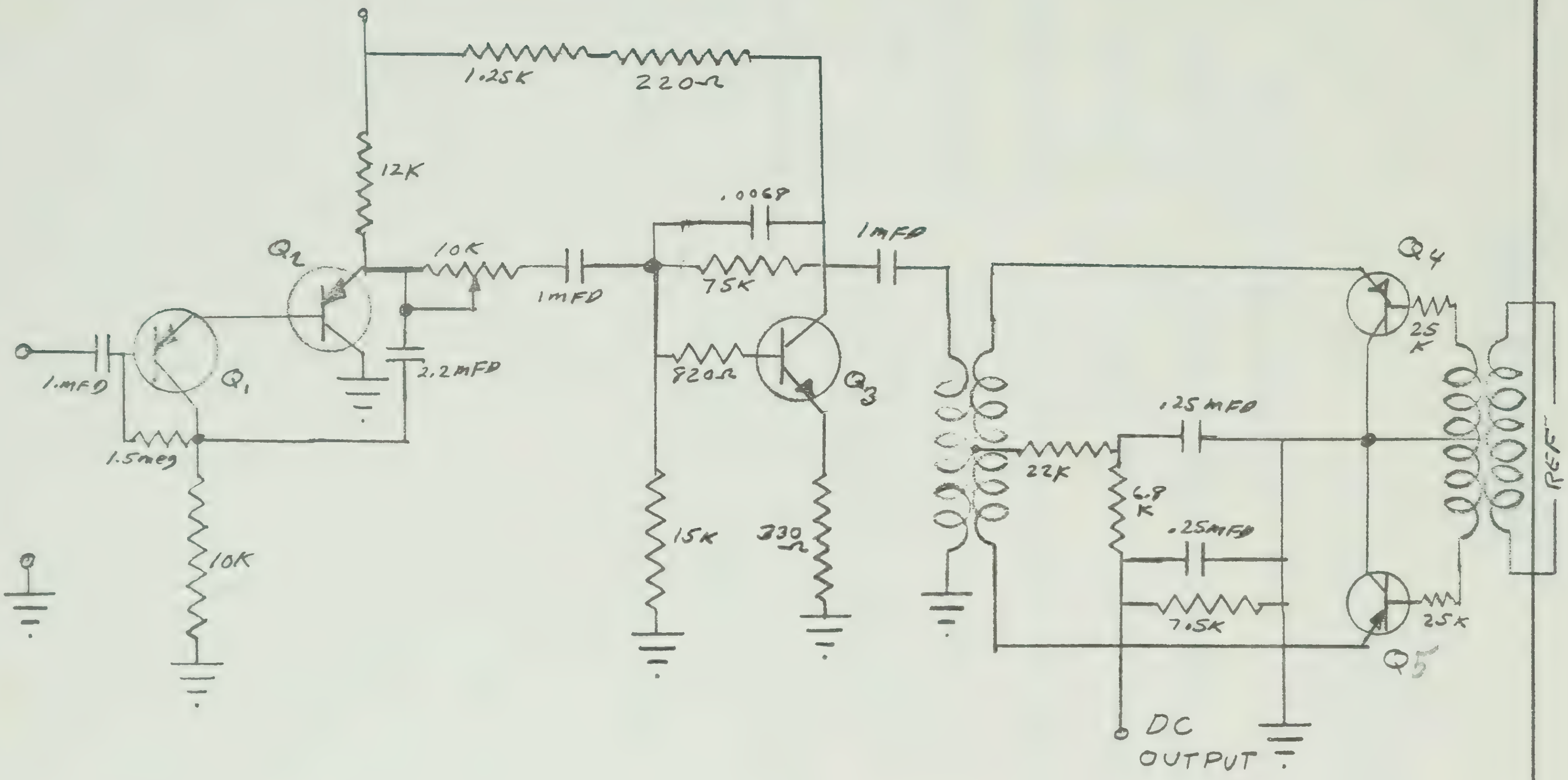
FREE Gyro



aCCELEROMETER Rate Gyro Simplified



SIMPLIFIED E TRANSformer Pickoff



PHASE SENSITIVE CONVERTER

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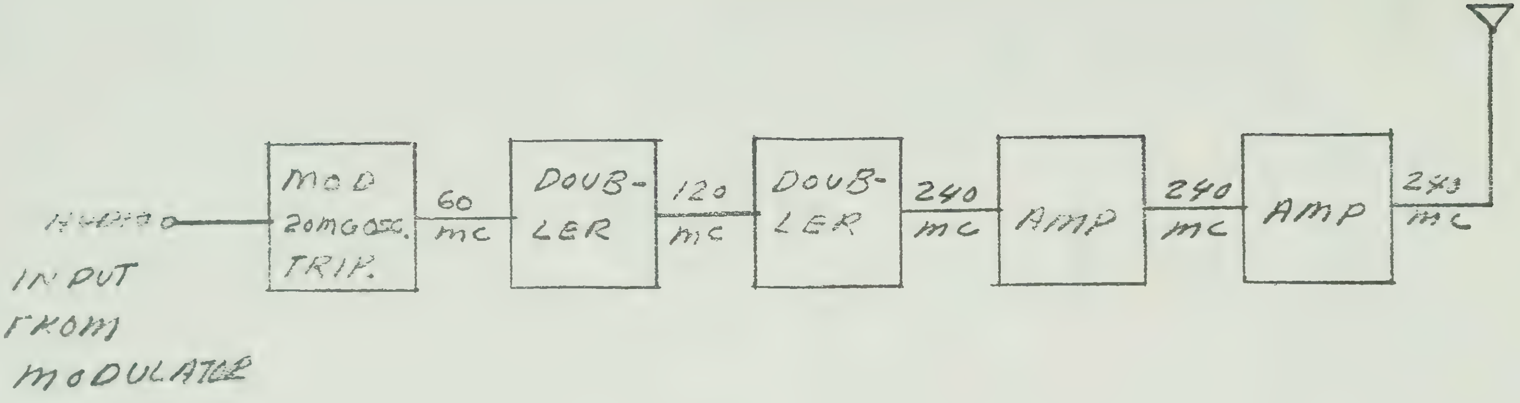
MISSILE SYSTEM DIVISION
XN NAVY TRAINING SCHOOL

Contract NOp-834
Course Notes SC-3-d

TOPIC: Transmitter Characteristics

Power Requirement:	6.3 volts ± 5% at 2.25 amperes 350 volts DC ±3.0% at 110 milliamperes
Modulation Input:	5.0 volts RMS ±5%, 0-100 KC, across 50 K ohms
Frequency Deviation:	±150 KC maximum (not to exceed ±125 KC in operation)
Frequency Stability:	±0.01% of center frequency
Power Output:	5 watts min. into 50 ohm resistive load; 4 watts min. into 50 ohm reactive load of 2:1 S W R
Frequency Distortion:	Less than 2.0% in range of 0 to 100 KC
Frequency Range:	Model TMR-1A 245-265 mc Model TMR-1B 225-245 mc
Spurious output:	60 db Min. for all unwanted freq.



TRANSMITTER - T/M

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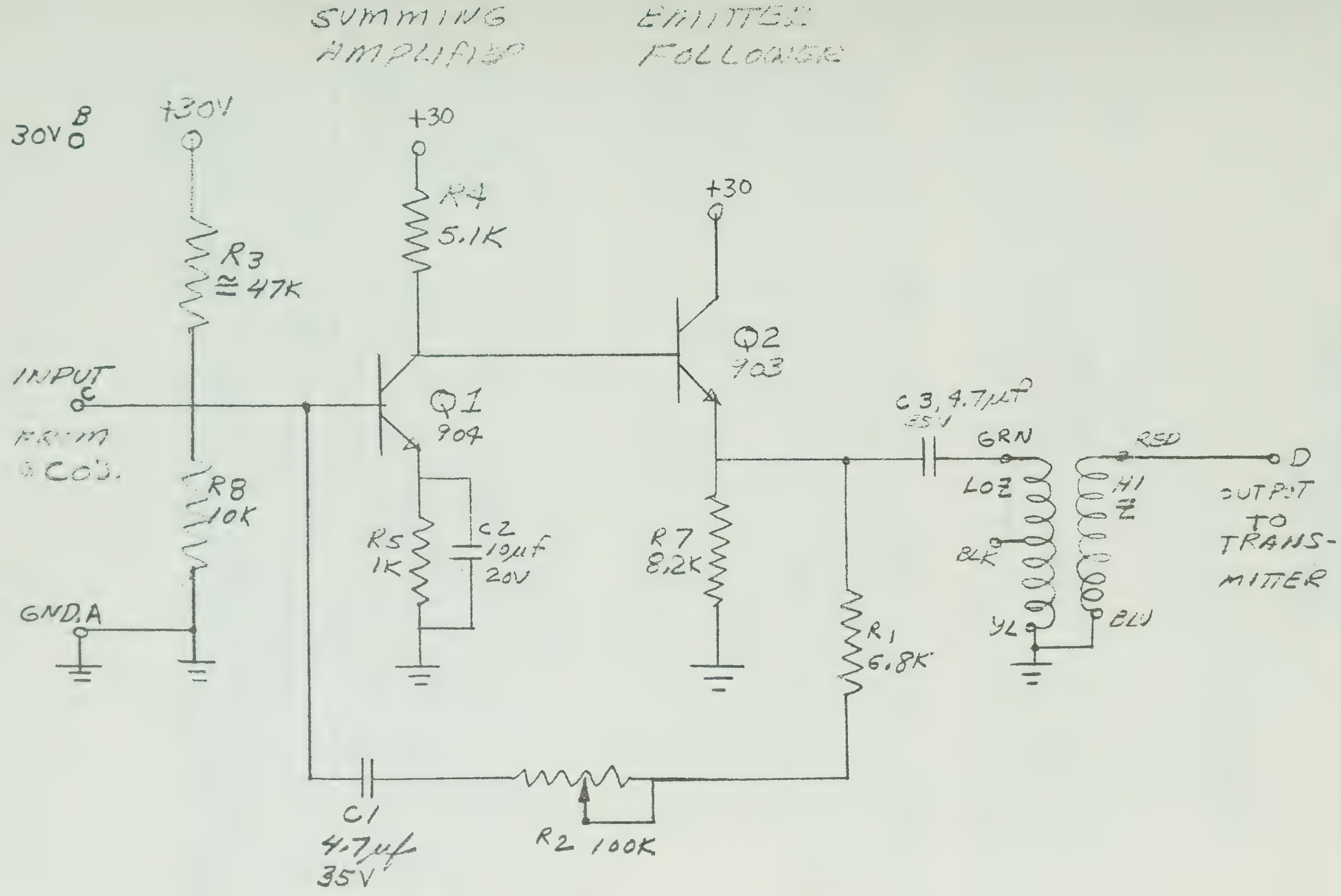
MISSILE SYSTEMS DIVISION
XN NAVY TRAINING SCHOOL

Contract NOp-834
Course Notes SC-3-c

TOPIC: Modulation Amplifier Characteristics

This component will serve to amplify the composite subcarrier signal to a level sufficient for modulation of the transmitter. Component performance shall conform to the following specifications:

Power Requirement:	28 VDC ± 10%
Gain:	200
Frequency Response:	±4 db maximum from that measured at 1 KC from 100 to 80 KC
Distortion:	0.5% maximum measured at 10 and 20 KC and including harmonies to 60 KC
Signal Output:	5 volts RMS maximum requirement, adjustable
Output Impedance:	1 K ohms maximum



MODULATION AMP. MOA-2

LOCKHEED AIRCRAFT CORPORATION
MISSILE SYSTEMS DIVISION
XN NAVY TRAINING SCHOOL

Contract NOp-834
Course Notes SC-3-g

TOPIC: In-Flight Calibration

As missile performance must be judged almost exclusively by data telemetered to the ground, the accuracy and reliability of the telemeter system is naturally of prime importance. Two different in-flight calibration schemes are utilized on the Polaris flight tests to either monitor, or to increase, the accuracy of the system.

On commutated channels, as shown in Figure 1, calibration is accomplished each rotation of the wiper, giving a minimum of 5 calibrations each second. The VCO is especially adapted to such calibration. A positive, a zero, and a minus voltage are applied to the three calibrate contacts. The amplitudes are adjusted to give an oscillator frequency shift of +42.5% of B.W., center frequency, and -42.5% of B.W., respectively. The figure also shows a simplified commutator record. Any change in either sensitivity or in drift of center frequency is automatically compensated for by the data reduction computer. There is no compensation (as only three points are checked) for linearity changes, but experience shows that such changes are relatively rare.

On continuous channels, shown in Figure 2 the calibration cycle interrupts the flow of data and such calibrations are restricted to a time close to some critical measurement, or to a time just following some severe environmental conditions (such as after launch or stage separation). The in-flight calibrate cycle again consists of a plus, zero and minus sequence and is initiated by a command signal from some internal event (like first separation). The in-flight calibrate unit is a timer and controls the operation of relays which (with the DLO) cause a known excitation of two different phases to shift the oscillator frequency 35% of B.W. As with the commutated data, there is automatic computor compensation for drift and sensitivity, but as the history between the two calibrate points is not known, the accuracy is improved only on those readings close to the calibration points. The prime purpose of continuous channel in-flight calibration, however, is to increase the confidence level by demonstrating that drift has been less than the tolerance expectations of the oscillator.

Figure 3 shows a representative Polaris flight and the in-flight calibration times are identified by arrows. The first calibration occurs just after first separation, or after approximately one minute of powered flight. The total time lapse until telemetry is lost is less than two minutes - a relatively short period of time for temperature or other drift effects to occur. During the majority of the flight, there is no telemetered data and a hundred miles is a reasonable estimate of the distance from the target when acquisition of telemetered data occurs. Calibration is then performed upon re-entry, and with less than one minute left before impact. There is thus a high order of confidence in the accuracy of the critical re-entry measurements.

The in-flight calibrate module consists of essentially a chain of timing circuits which sequentially operate relays to provide calibration signals for the VCO's, the DLO's and the PSC's. It also provides a delayed trigger, permitting the initiation of the calibrate cycle to be delayed for 3.5 seconds after first separation. Large transients, which may be caused by such separation, are thur allowed to damp out before calibration begins.

The in-flight calibrate unit provides an in-flight calibration for continuous channels, similar to that provided on commutated channels. Its main purpose is to assure that telemeter components have stayed within their tolerance specifications. The VCO's, the DLO's and the PSC's are all checked for zero drift and sensitivity changes.

Simplicity, ruggedness, reliability, and low manufacturing cost have been emphasized in the design of this unit. All transistors are operated as switches and timing is controlled by RC networks. The voltage at which transistor switching occurs (on the condenser charge curve) is regulated by Zenner diodes. As gross timing errors cannot be introduced by voltage fluctuations, specifications can be met without temperature compensating the RC networks. Thus, no selection of parts or transistors is necessary.

SPECIFICATIONS

Table I list the important electrical and mechanical characteristics of the in-flight calibrate unit.

TABLE I

SPECIFICATIONS FOR THE IN-FLIGHT CALIBRATE UNIT

Schematic:	1039261
Acceptance Test Specifications:	1700495
Manufacturing Test Procedure:	

MECHANICAL

Size:	
Weight:	

TRANSISTOR COMPLEMENT

Q1	2N335
Q3, Q6, Q9, Q10, Q13	2N338
Q2, Q5, Q8, Q12	2N495
Q4, Q7, Q11	2N497

ELECTRICAL

Power Requirements:	30 V DC
	Calibrate - 100 ma. max.
	Standby 12 ma. max.
Calibrate Cycle:	Cal Z, 3.3 +0.6 - 0.3 seconds
	Cal Plus, 0.5 ± 0.15 seconds
	Cal Minus, 0.5 ± 0.15 seconds
Trigger (Command):	27 ± 3V DC, 50 MS duration
Trigger (Delay):	27 ± 3V DC, 50 MS duration
Retrigger (either):	Circuit capable of operating within 4.0 seconds after completion of a normal double cycle.
Delay Timer:	
Environmental Specs:	

Figure 4 is a detailed block diagram of the in-flight calibrate unit. It contains five similar timers, three of which operate the control relays and cycle the calibration voltages to the subcarrier oscillators. Examination of the top waveforms in figure 5 shows that the Zero Cal control relay is on for approximately 0.5 seconds per cycle. A typical calibration signal to the VCO's is also shown for reference. The period between the Plus and Minus calibrations represents the 0.5 seconds that the interval timer is on. The logic of the in-flight calibrate circuit is very simple. Zero alone is zero. Zero and plus is plus. Zero and minus is minus.

The Zero Cal timer, as shown in the block diagram, can be started by a trigger from the Command No Delay input, or by a trigger from the 4 second delay timer. When the Zero Cal timer is turned on, it supplies an "on" ground to both the Plus Cal Timer and the Minus Cal timer. It also supplies a positive trigger to the Plus Cal Timer. Both control relays thus go on together. The Plus Cal control delay releases after 0.5 seconds and a trigger is sent to the Interval timer. After an additional 0.5 seconds, or at t_0 +1, the Interval timer triggers the Minus Cal timer and this relay operates. At the end of its 0.5 second "on" time the Minus Cal relay releases and a trigger is sent to the Plus Cal timer at t_0+2. This trigger starts a second sequence of calibration cycles.

The Plus Cal, Interval, and Minus Cal timers thus form a loop which will continue to sequence until the Zero Cal timer goes off, opening the "on" ground and shutting them off. The bottom group of waveforms in figure 5 shows the entire sequence of trigger pulses.

DETAILED CIRCUIT ANALYSIS

Figure 6 is a schematic representation of the basic circuit used for the Delay Timer, the Zero Cal timer, the Plus Cal timer and the Minus Cal timer. Except for minor differences in some resistors, the circuits are almost identical. Major differences arise in the load resistor R3, which is replaced by relays in all but the delay timer, and the time constant formed by R4 and C2. The transistors used in each timer have been identified for reference.

In its stable state, the base of Q1 is held negative by the bias applied through R2, and Q1 is "off". The collector voltage is +28 V and C2 is charged to only 4 V. The other side of the condenser is at approximately +24 V, set by the divider formed by R4 and R6 through CR4. Q2 is held "on" by this voltage and as it is an emitter-follower, a +24 V also appears across R7. Q3 is a PNP transistor whose emitter is held at +18 volts by the Zenner diodes CR5 and CR6. As the base of Q3 is more positive than its emitter, it is "off". Its collector, which is tied to the base of Q1, is at approximately -0.5 V.

When a positive trigger is applied to the base of Q1, it is turned "on" and the collector voltage drops to approximately 1 V. C2 now starts charging and the charging current drops the voltage at the lower end of R2 to 1 V. As the base of Q2 is near ground potential while the emitter is held positive, Q2 is now cut off. The voltage across R7 cannot, of course, fall below +17.5 V. At this point Q3 turns on as its base is now less positive than its emitter. As long as Q3 stays on, it will also hold the base of Q1 at approximately +17 volts. Q1 is thus held "on" until C2 charges up to a +18 volts. At this point the base of Q2 becomes more positive than its emitter (which is held at 17.5) and Q2 turns "on" again. The voltage across R7 now rises to +24, cutting off Q3. The collector of Q3 and the base of Q1 now falls to a -0.5 volts and Q1 returns to its normally "off" position. The length of time it takes C2 to charge to a +18 V is, of course, a function of the time constant formed by C2 and R4.

The same analysis would apply to the Zero Cal Timer, with the exception that R3 is now a relay, which operates whenever Q4 goes on. As the Zero Cal Timer must go "on" at the end of the delay interval, the trigger is taken from the collector (Point a) of Q1 which goes positive at the end of the delay interval. This positive trigger is applied to Q4 via point (b).

The triggering requirement is slightly different for the Plus Cal timer. The Plus Cal relay must turn on at the same time as the Zero Cal relay. The + trigger is thus obtained from point (c) or the collector of Q6 in the Zero Cal timer and RC coupled to point (b) or the base of Q7 in the Plus Cal timer.

Figure 7 shows the method by which the Zero Cal timer supplies the 3 second "on" ground to the Plus and Minus Cal timers. This also provides a fail-safe feature and component failure in the In-flight Calibrate unit will cause loss of calibration rather than loss of data. The emitters of both Q7 and Q11 depend on Q4 for their ground return. The Plus and Minus control relays can therefore only operate if Q4 in the Zero Cal timer is "on."

The Interval Timer, shown in Figure 8, provides a 0.5 second period of zero calibration, during which neither the Plus or Minus Cal relays are on. Q10 is normally "on" as its base is held positive by R25, CR14 and R27. When Q7 is turned on during the Plus Cal interval, its collector drops to 1 V, turning on CR15 and dropping the voltage at the junction of C9 and CR14 to a low positive value. C9 charges rapidly to 28V. As the base of Q10 is now negative in respect to its emitter, which is held at +18 V, Q10 is turned off. At the end of the Plus Cal cycle, the collector of Q7 returns to a +28 volts and CR15 is open. The voltage at the junction of C9 and CR14 cannot rise to +28 volts until C9 discharges through R25.

When the voltage at the junction rises to +18 volts, Q10 turns "on" again. At this time a -10 V trigger is coupled through C10 to the Minus Cal Timer.

As the trigger is negative, it must be applied to one of the "on" transistors in the Minus Cal Timer. It is, therefore, coupled to point (d) or the base of Q12 (see Figure 6). The action of the Minus Cal timer, is then identical to the timers previously discussed. At the end of the Minus Cal interval, a positive trigger is taken from point (a) and RC coupled back to the Plus Cal Timer.

CALIBRATION CIRCUITS

Figure 9 is a simplified block diagram of the system calibration circuit. It represents a single telemeter link, each link contains its own in-flight Calibrate unit. Some liberties have been taken with the logic of the circuits, for the sake of clarity. The In-flight Calibrate unit sequences the control relays, as previously discussed. These relays, in turn, supply 28 volts to sequence the calibrate relays. The calibrate relays, in turn, do the actual cycling of calibration signals. Only one each of the three calibrate relays are shown in the block diagram. A large number of relays (which varies somewhat in the different links) are actually paralleled. The use of separate control relays, and the paralleling of many calibrate relays has two distinct advantages.

1) It increases system flexibility.

2) It enables relay contacts to be placed closer to the specific units being calibrated and, thus, avoids long lead runs.

The calibrate relays disconnect the inputs to the VCO's and the PSC's from their transducers and connect them instead to reference voltage sources. For purposes of clarity only, one PSC, VCO or DLO has been shown. There are, of course, many such units in each link - all of which are simultaneously calibrated. By application of Plus, Zero and Minus calibrate voltages to the input of the VCO, the output frequency is shifted from +42.5% of B.W., through center frequency, to -42.5% of B.W. The input to the PSC is a 1 V rms signal of 800 cps reference phase, which is first grounded, and then reversed in phase. No external excitation is applied to the DLO as transducer excitation must be at oscillator frequency and frequency modulation is achieved through internal feedback. Internal switching by semiconductor diodes provides internal feedback for calibration purposes. The calibrate relays simply supply the -30 V to the DLO to operate these internal switches.

Figure 9 shows the detailed circuits for the VCO calibration. The reference voltage is supplied by mercury cells whose output, after aging, varies as little as 0.15% over a 300 hour period. Three different calibration sequences are supplied to terminals on the Mod Terminal Board:

	Plus	Zero	Minus
(a)	+0.75	0	-0.75
(b)	+1.50	+0.75	0
(c)	0	-0.75	-1.50

The range of each VCO must be set according to the particular transducer with which it is used. If, for example, a pressure transducer is monitoring 0 ± 15 PSIA, its output will probably be 0 ± 0.75 volts and center frequency will be with 0 volts input to the VCO. A 0 - 5000 PSI transducer, however, might have an output of 0 to +1.50 volts. The VCO must now be set so that center frequency occurs with +0.75 V input. Once the VCO has been set, the calibration signals must obviously match the transducer outputs. Selection of the proper calibration signals is accomplished by jumpering to the proper terminal.

The details of the PSC calibration are shown in figure 9. The same 800 cps reference signal which excites the gyro pickups is used for calibration. The center-tapped transformer provides phase-splitting and the resistors adjust the amplitude of the signal to approximately 1 V rms. The In-flight calibrate and control relays are also shown for reference.

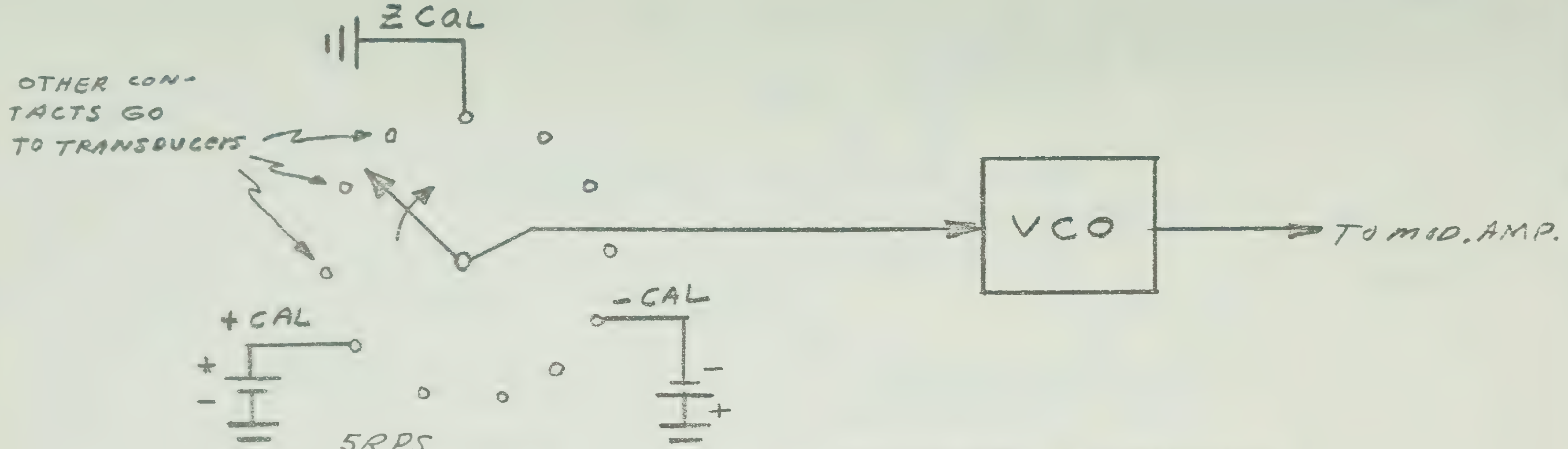
Two different calibration circuits are shown in figure 10, corresponding to the two different DLO's. The DLO-1, 2 has no provision for plus or minus calibrate (pins L and K being open). The Cal Zero is removed during the Plus and Minus Cal periods. The DLO-1A, -2A has Plus and Minus calibration capability. The Zero Cal is on for the full cycle. The +28 V DC through the 2 meg resistors provides a small + voltage for back-bias on the semiconductor switches inside the DLO, holding them off. During the calibrate cycle - 30 V inputs are applied, turning these switches on.

COMMAND SIGNALS

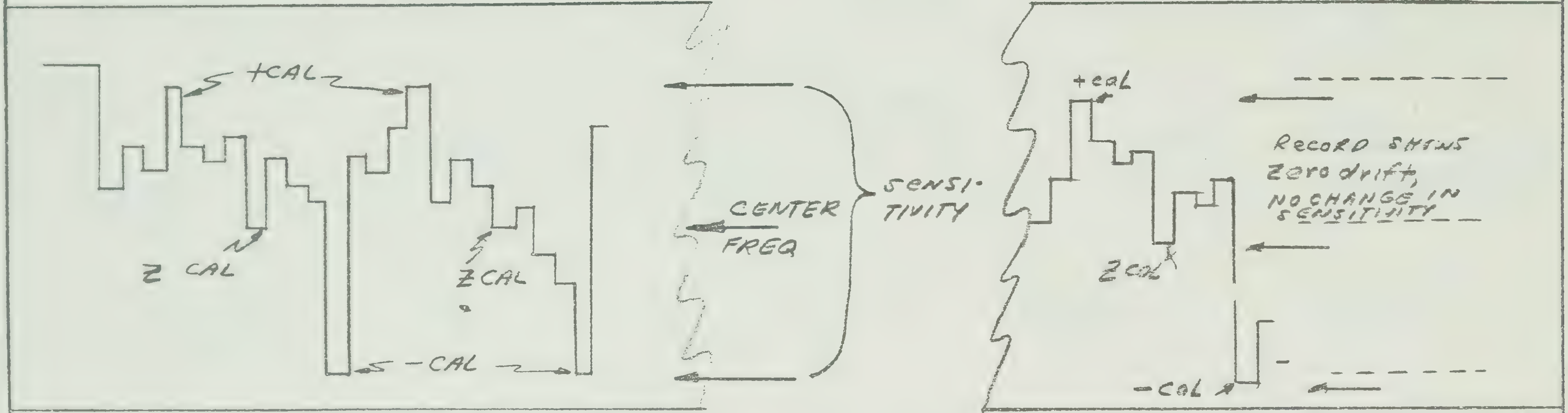
The Command Signal wiring for the in-flight calibration is shown in the simplified block diagram of figure 11. Prior to launch, and during checkout, a +28 V DC is supplied from the block house through the Power Transfer switch to energize the checkout relay. By then pushing the Delay switch, it is possible to cycle the entire in-flight calibration sequence through the Delay Command input.

The In-flight Calibrate units, in all three links, can also be operated by a command signal from the block house, or the In-flight calibrate can be bypassed and the calibrate relays operated directly (see also figure 9). At launch, power to energize the checkout relay is supplied from the vehicle instead of the blockhouse. At first separation, however, the relay ground connection is broken, allowing the relay to release. This applies a +28 V DC from vehicle power to the Delay command input, cycling the In-flight Calibrate unit. During re-entry the de-acceleration G switch initiates the timer and 15 seconds later a +28 V is applied to the Command input of the In-Flight Calibrate units in the first two links.

SC-39



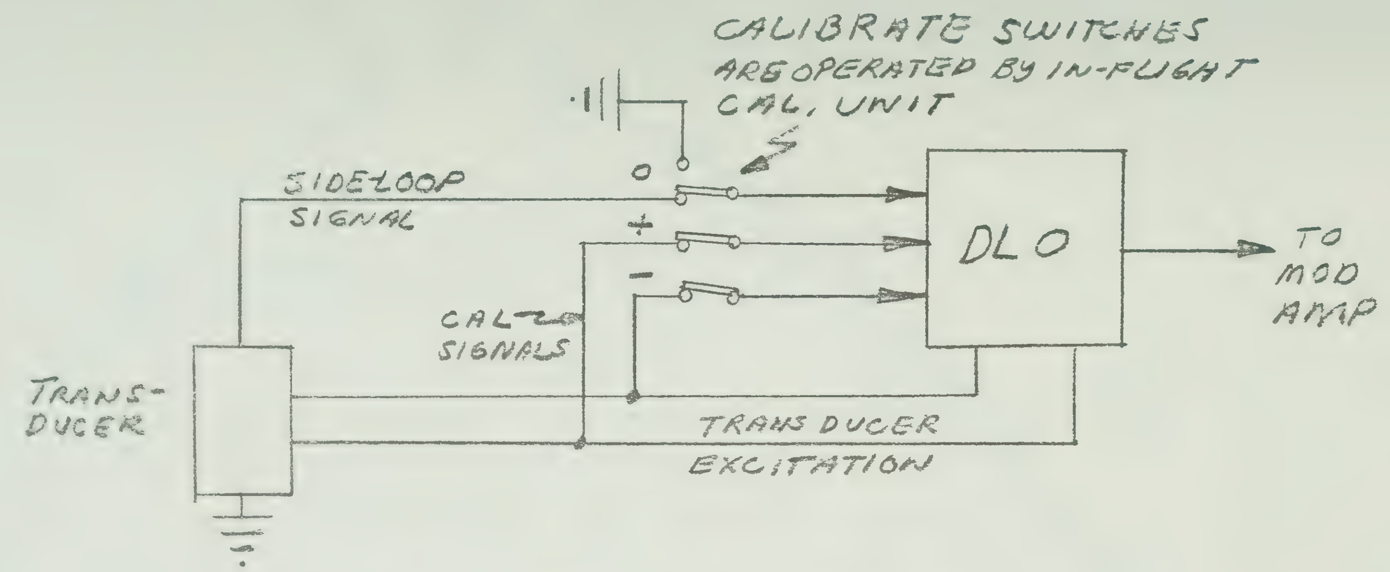
CALIBRATE TECHNIQUE



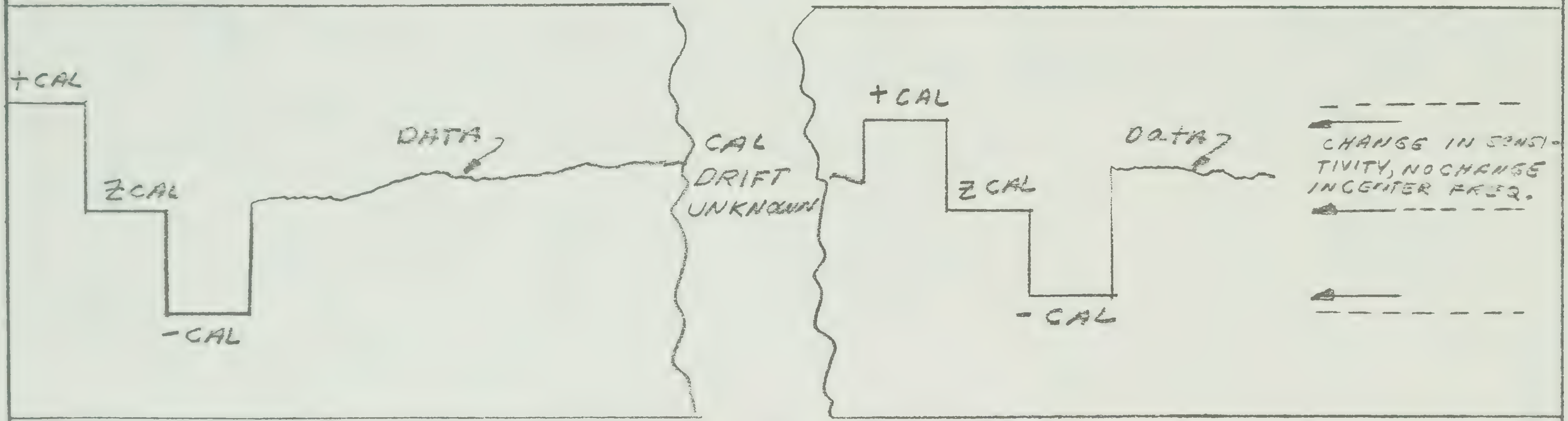
DATA RECORD

CALIBRATION OF COMMUTATED OSCILLATOR FIG-1

SC-3-9



CALIBRATION TECHNIQUE



DATA RECORD

CALIBRATION OF CONTINUOUS CHANNEL FIG-2

SC-3-9

SC-3-9

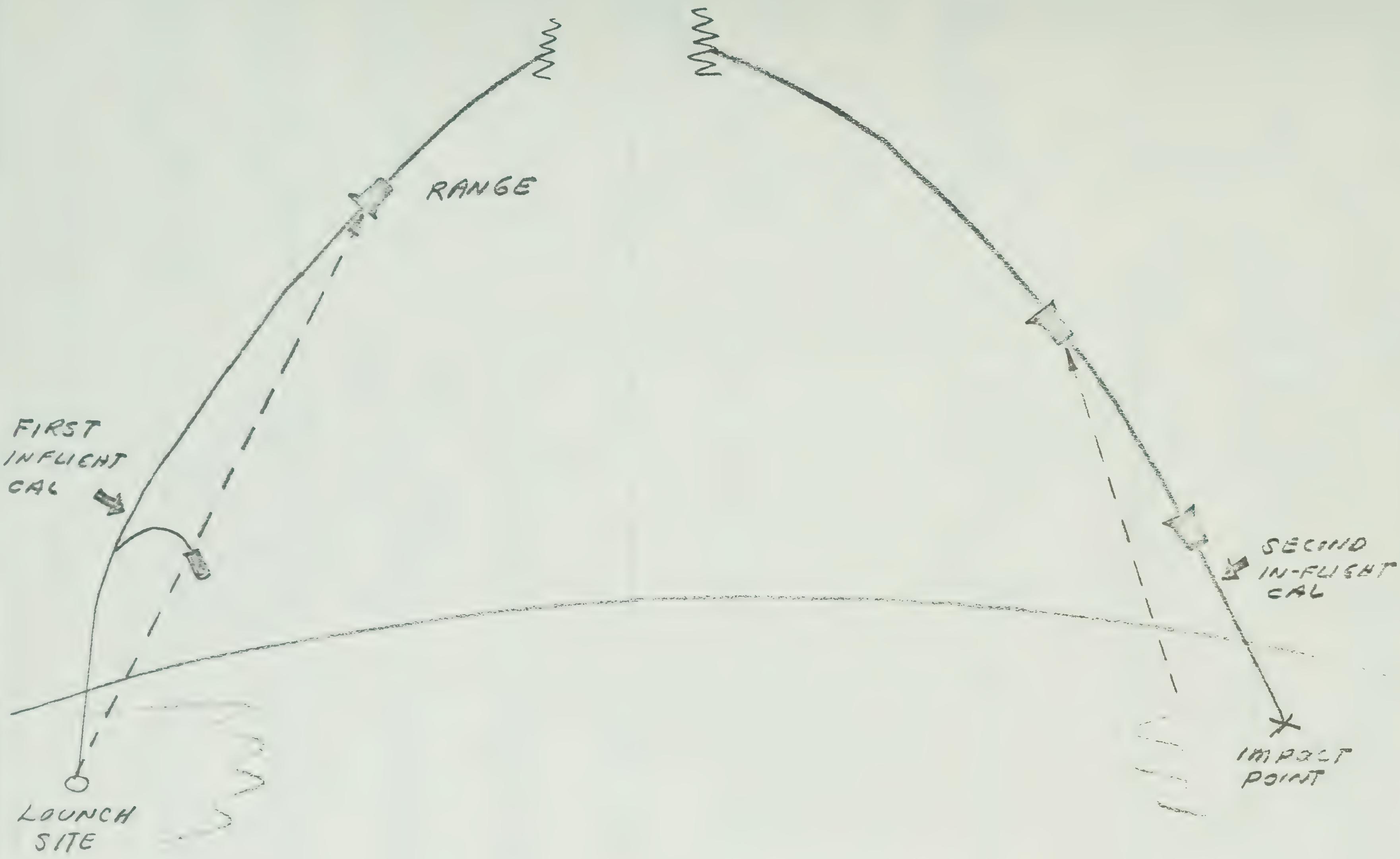
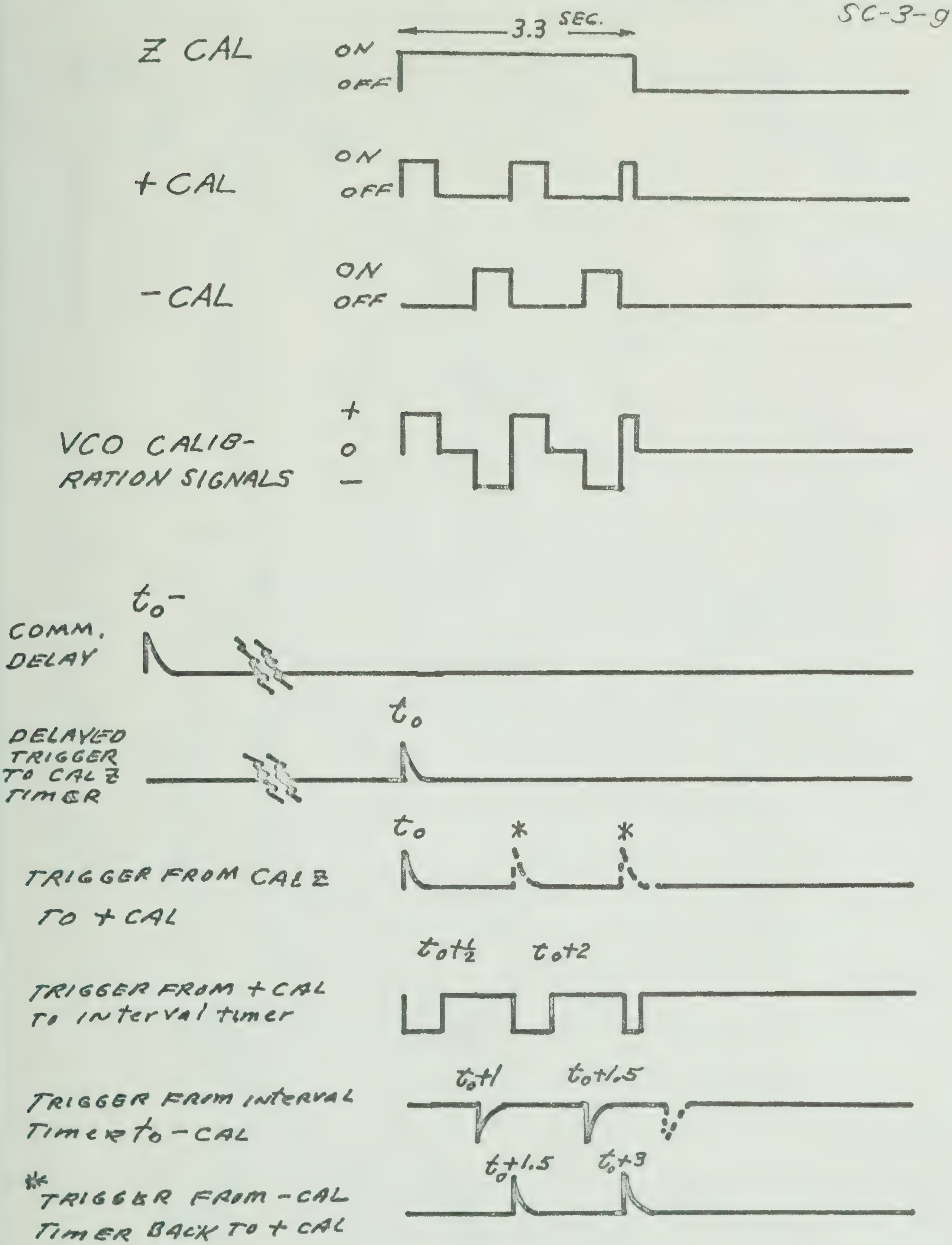
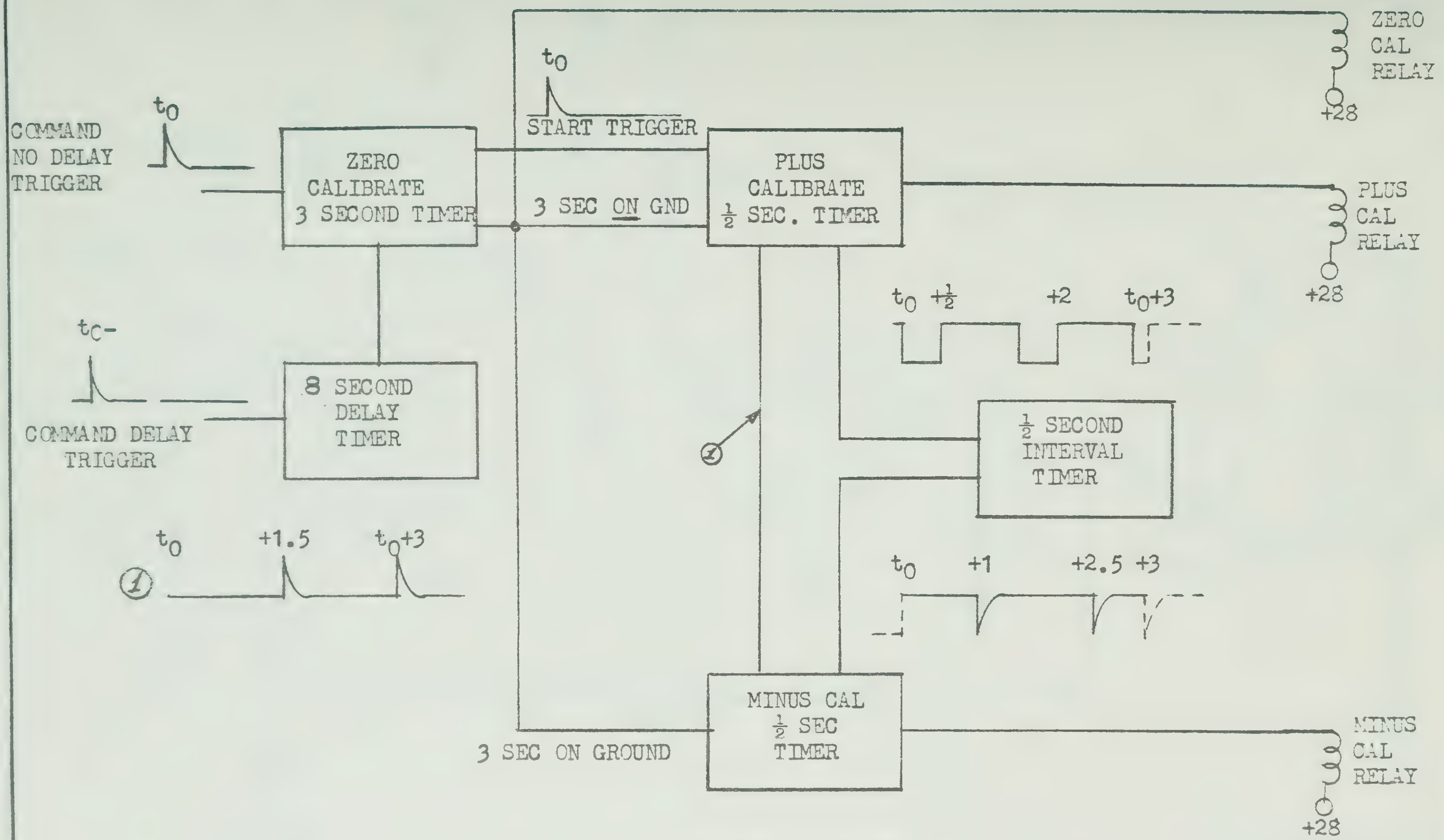

RANGE
FIRST
INFLIGHT
CAL
LOUNCH
SITE
SECOND
IN-FLIGHT
CAL
IMPACT
POINT

FIG-3


3.3 SEC.
Z CAL
ON
OFF
+ CAL
ON
OFF
− CAL
ON
OFF
VCO CALIBRATION SIGNALS
+
0
−
t_0-
COMM. DELAY
t_0
DELAYED TRIGGER TO CAL Z TIMER
t_0
*
*
TRIGGER FROM CAL Z TO + CAL
$t_0+\frac{1}{2}$
t_0+2
TRIGGER FROM + CAL TO INTERVAL TIMER
t_0+1
$t_0+1.5$
TRIGGER FROM INTERVAL TIMER TO − CAL
$t_0+1.5$
t_0+3
*TRIGGER FROM − CAL TIMER BACK TO + CAL

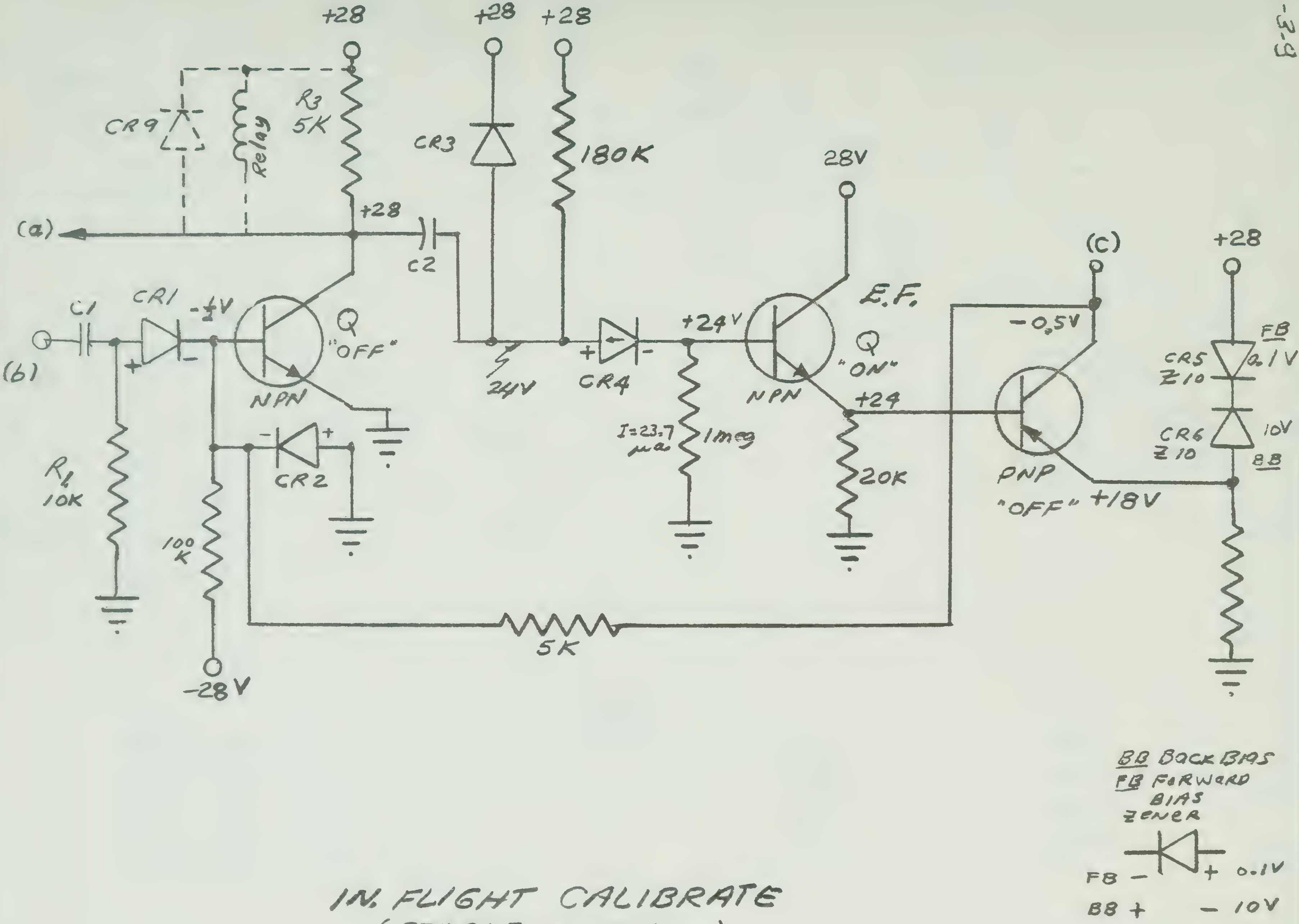
TRIGGER WAVEFORMS Fig-5



IN-FLIGHT CALIBRATE

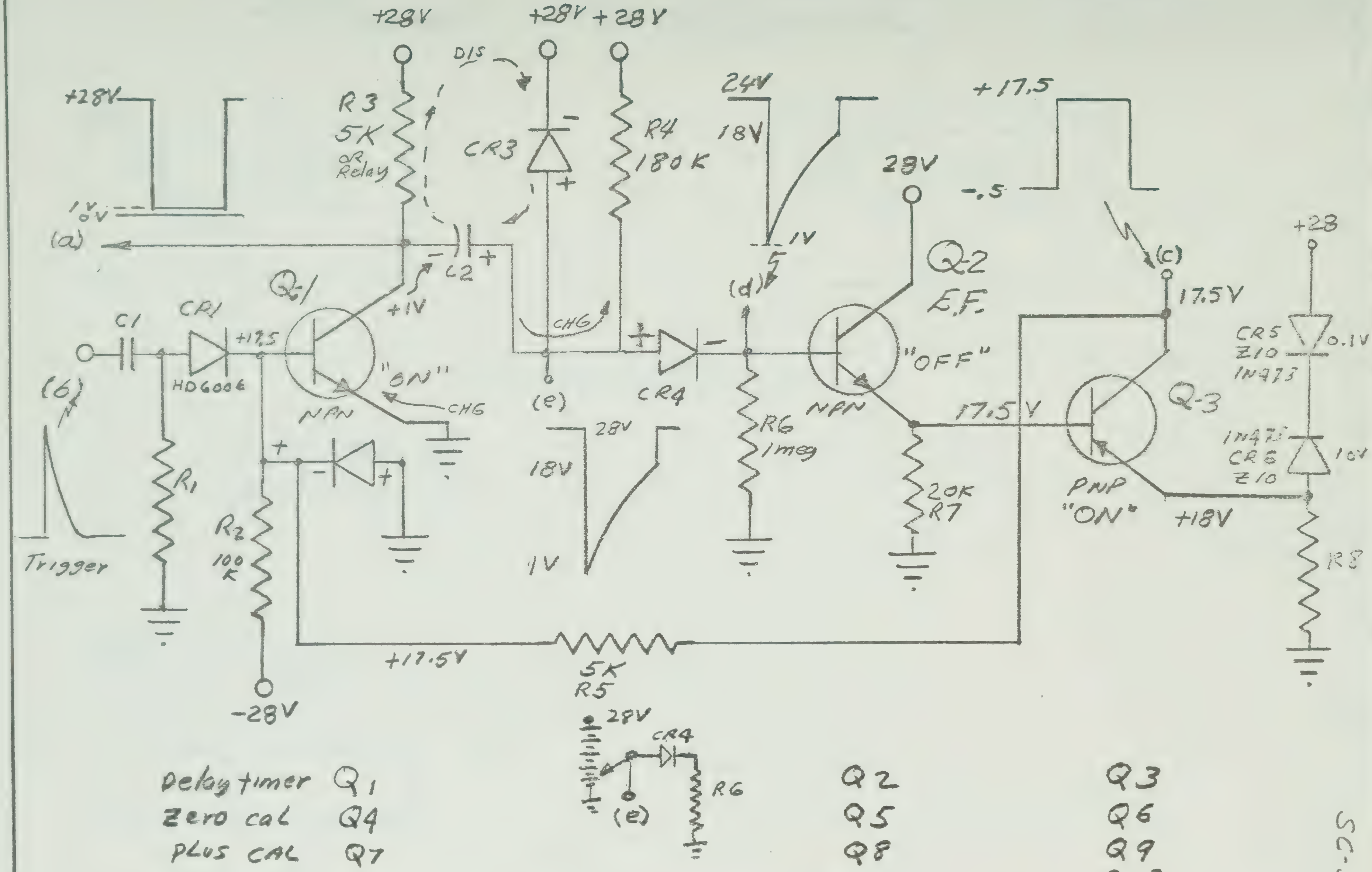
SC-3-9

SC-3-9



IN. FLIGHT CALIBRATE
(STABLE CONDITION)

SC-3-9



Delay timer	Q1	Q2	Q3
Zero cal	Q4	Q5	Q6
Plus cal	Q7	Q8	Q9
Minus cal	Q11	Q12	Q13

IN FLIGHT CALIBRATE

(TRIGGER APPLIED)

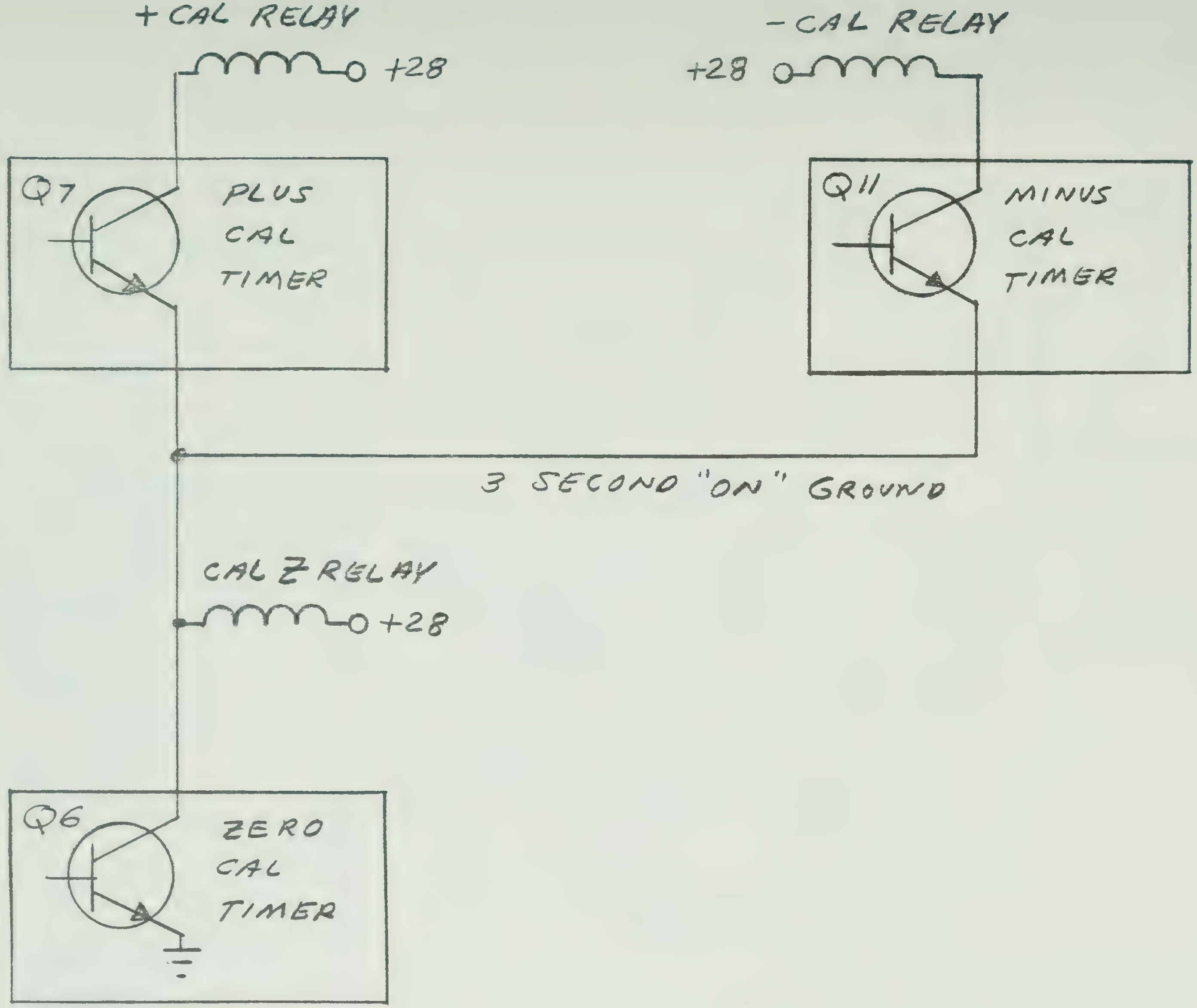

+CAL RELAY
+28
-CAL RELAY
+28
Q7
PLUS
CAL
TIMER
Q11
MINUS
CAL
TIMER
3 SECOND "ON" GROUND
CAL Z RELAY
+28
Q6
ZERO
CAL
TIMER

FIG-7

SC-3-9

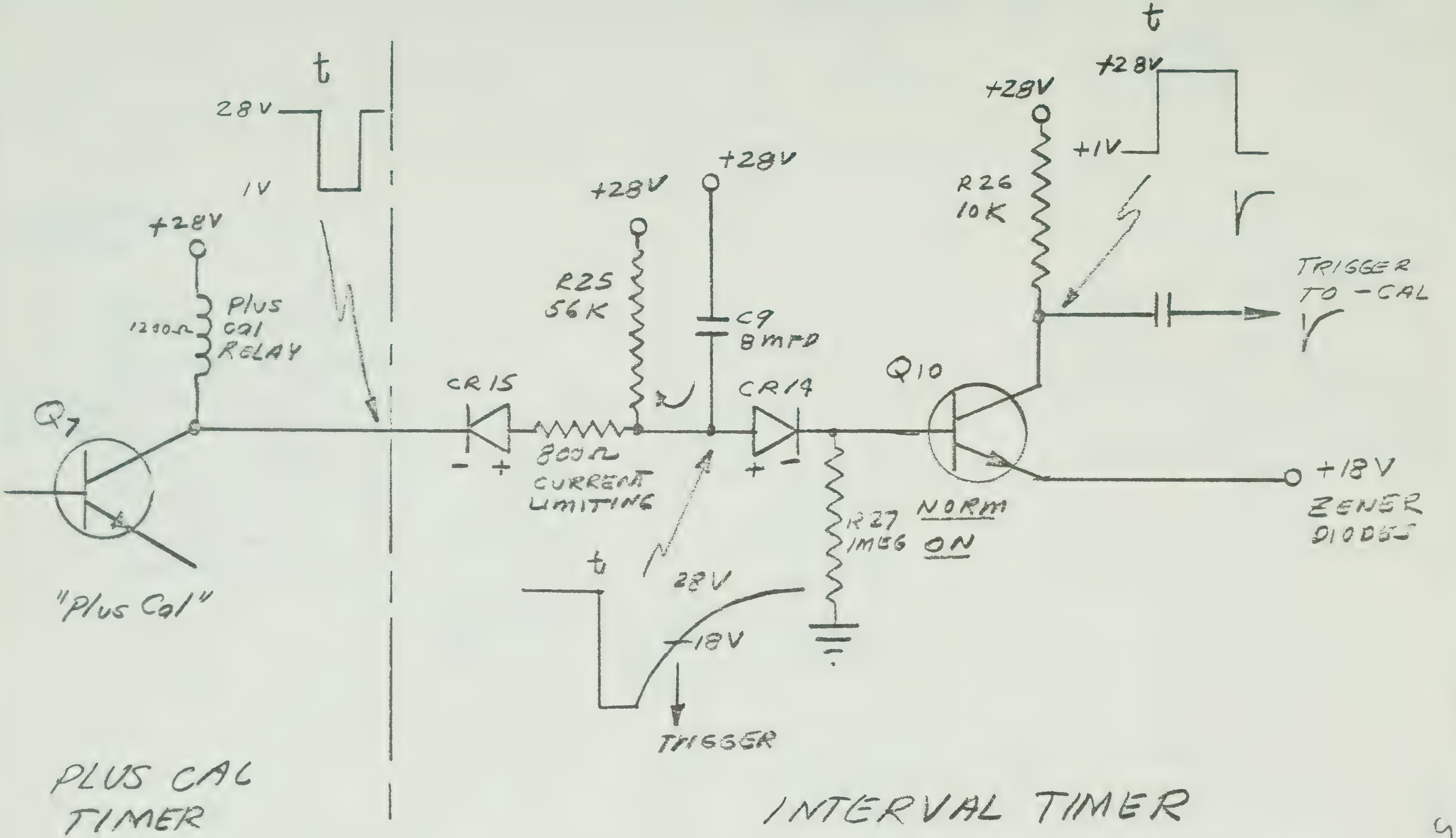
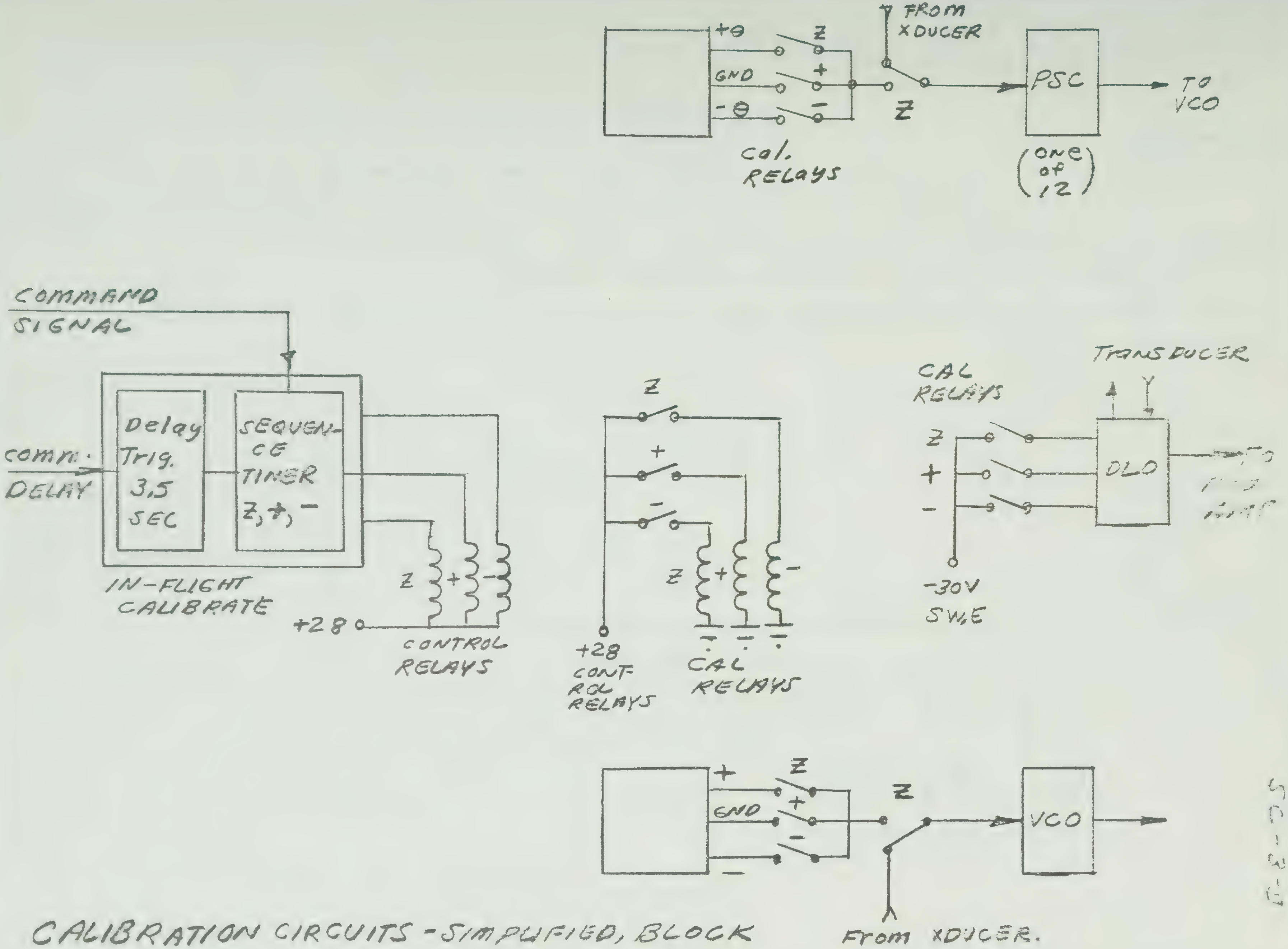

t
28V
1V
+28V
Plus
cal
RELAY
1200Ω
Q7
"Plus Cal"
PLUS CAL
TIMER
+28V
R25
56K
800Ω
CURRENT
LIMITING
CR15
- +
+28V
C9
8MFD
CR14
+ -
t
28V
18V
TRIGGER
R27
1MEG
INTERVAL TIMER
Q10
NORM
ON
+28V
R26
10K
t
+28V
+1V
TRIGGER
TO -CAL
+18V
ZENER
DIODES

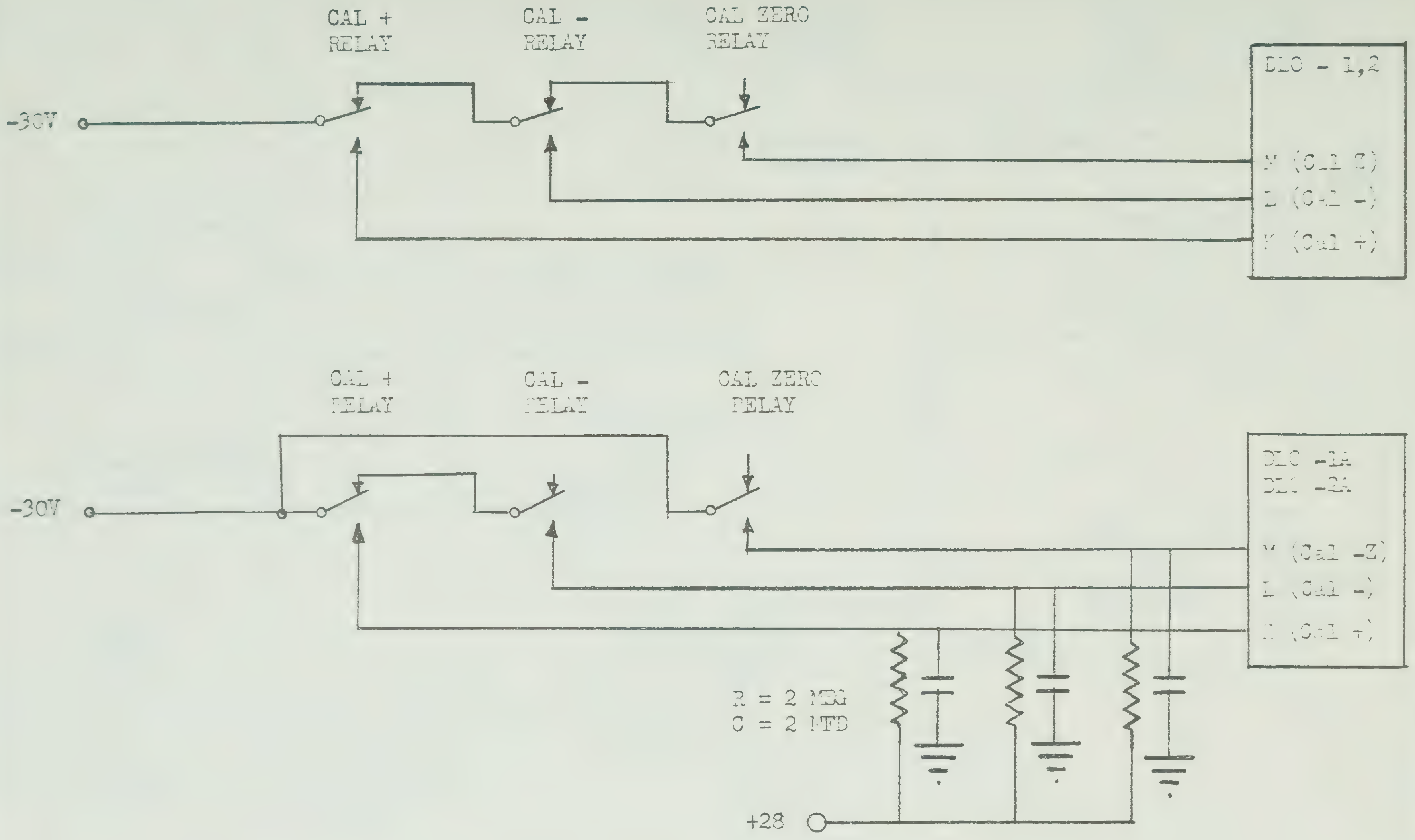
FIG-8

5C-3-9

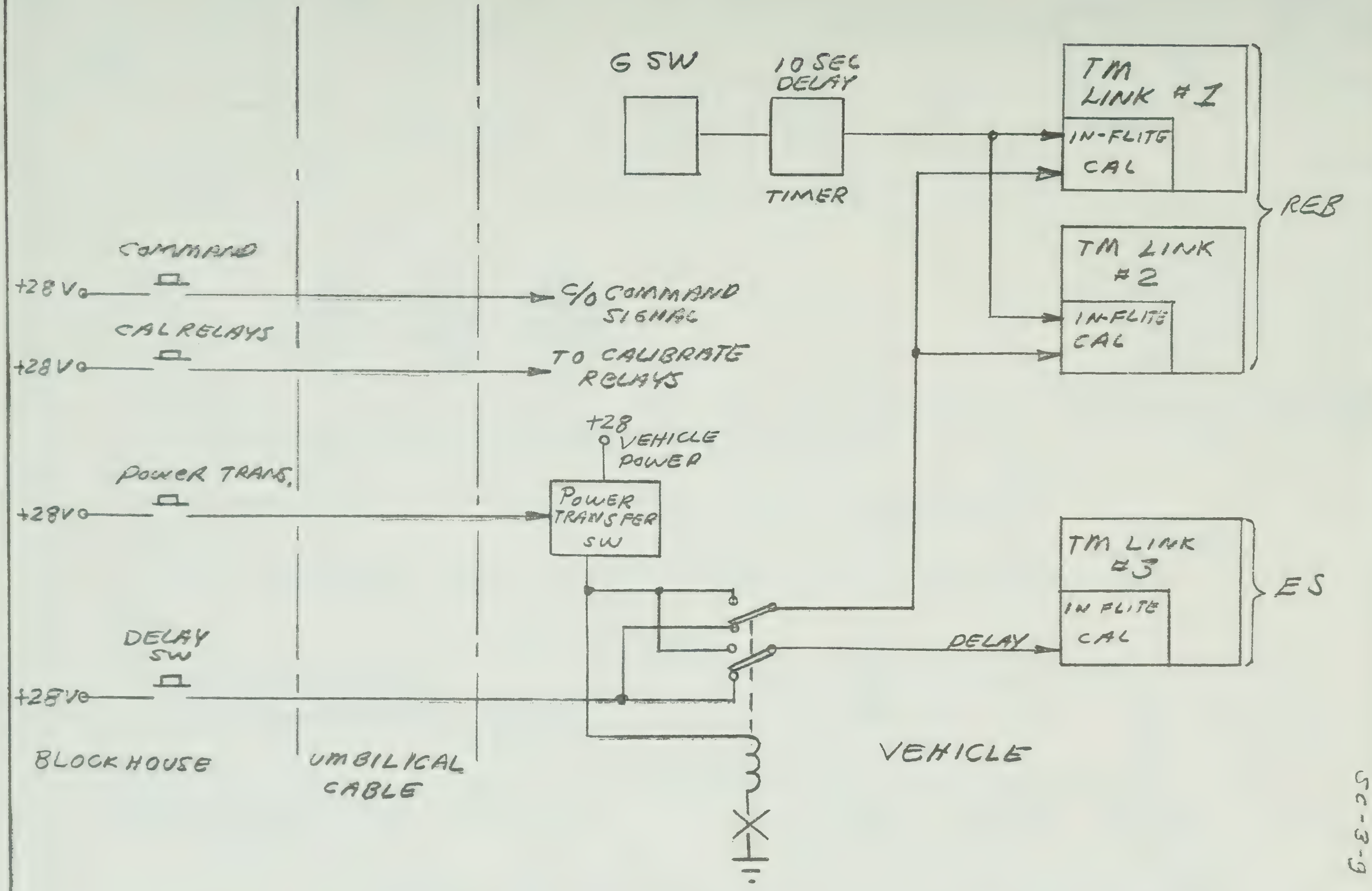


CALIBRATION CIRCUITS - SIMPLIFIED, BLOCK
Fig 9

SC-3-9

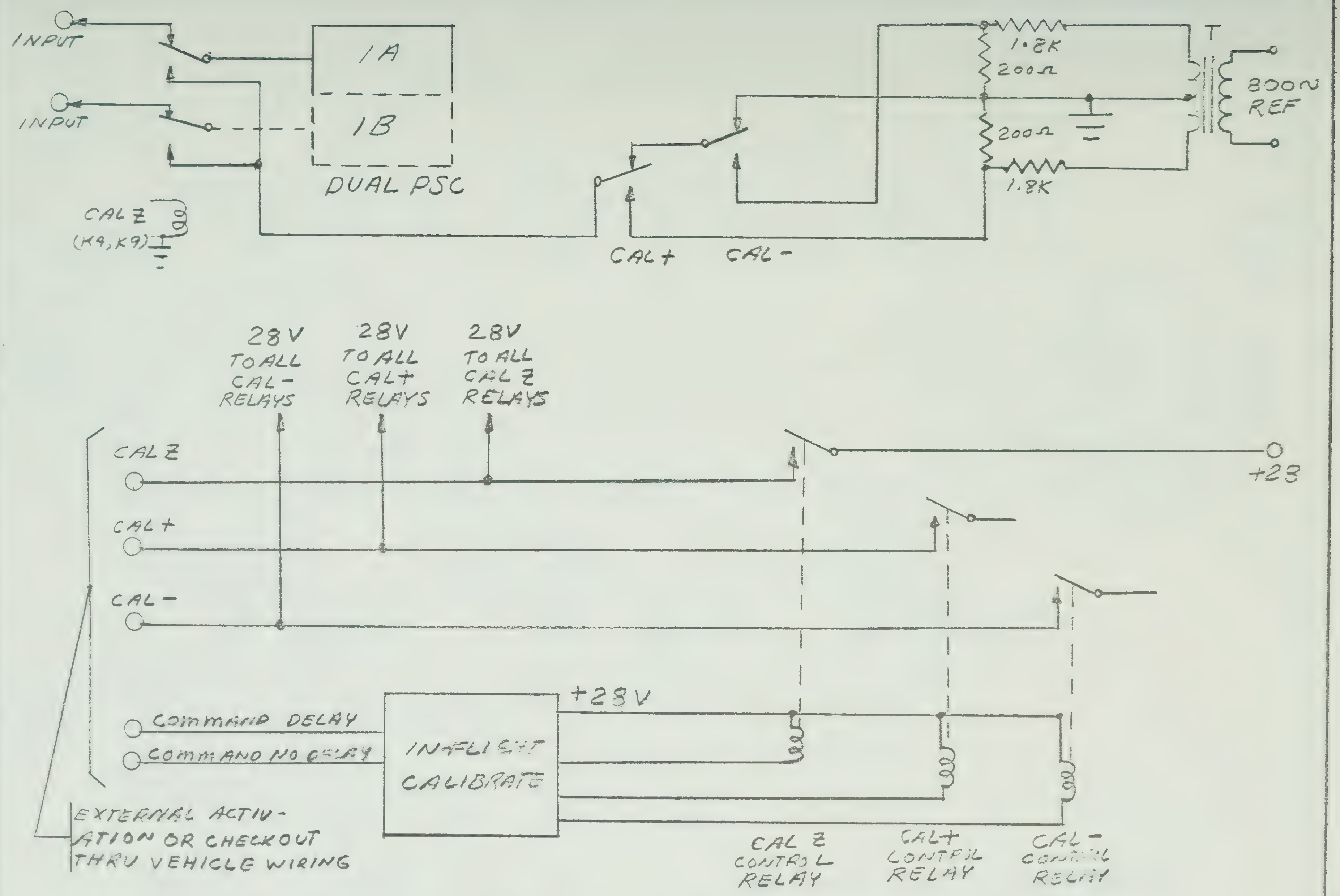


IN-FLIGHT CALIBRATE FOR DLO

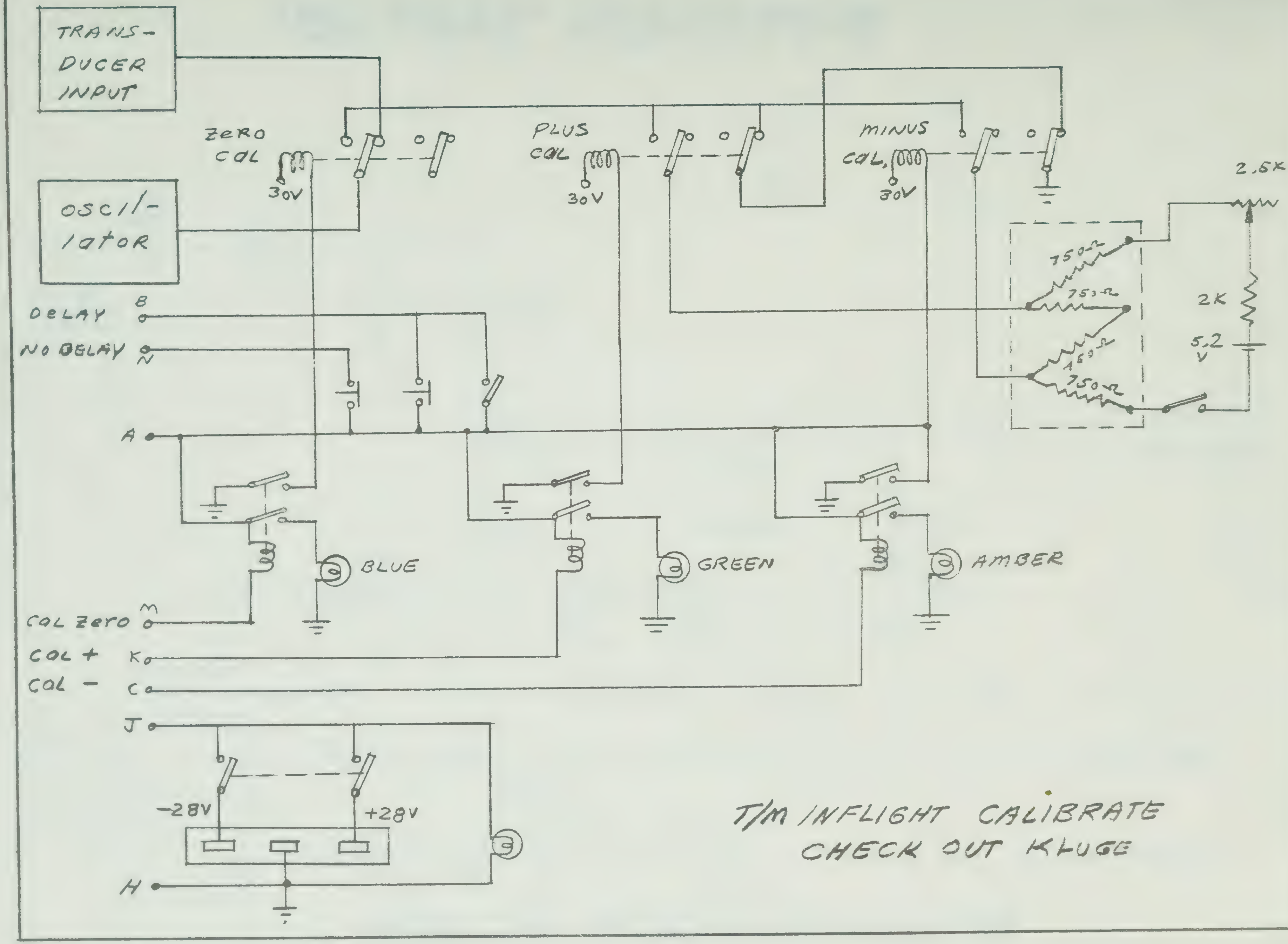


SC-3-9

COMMAND SYSTEM FOR IN-FLIGHT FIG-11



PSC CALIBRATION & IN-FLIGHT CALIBRATE
CONTROL RELAY CIRCUITRY

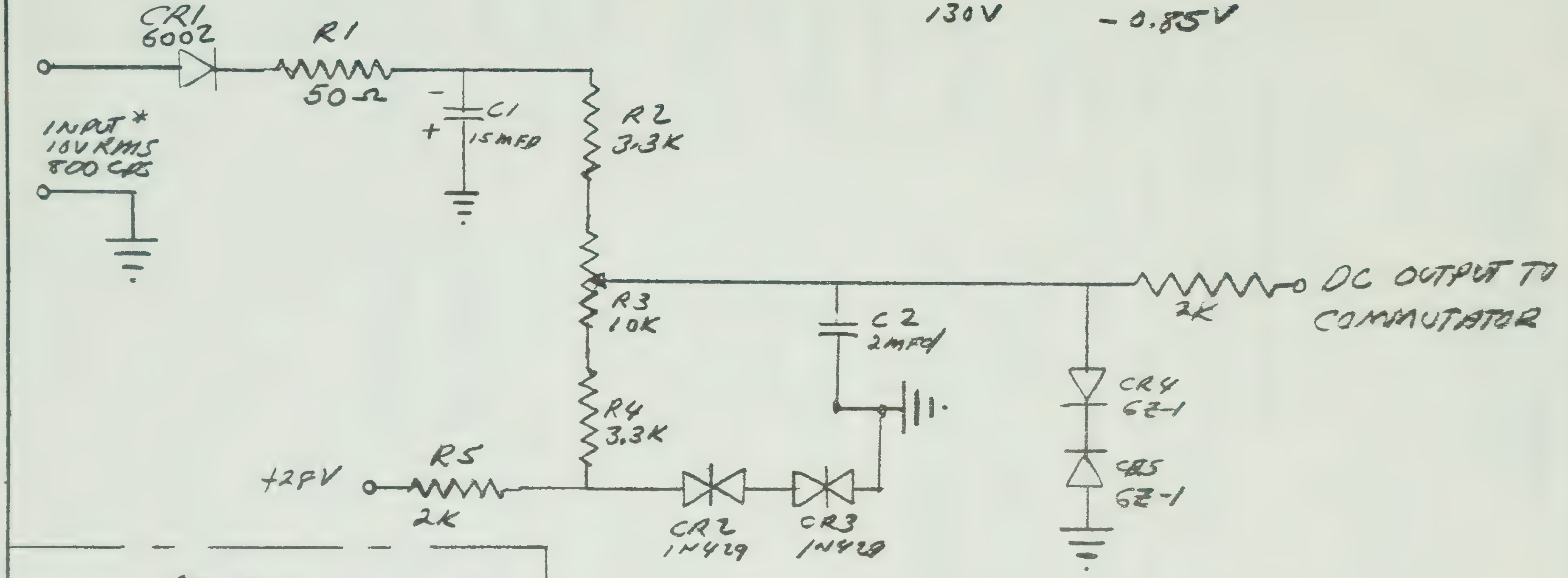


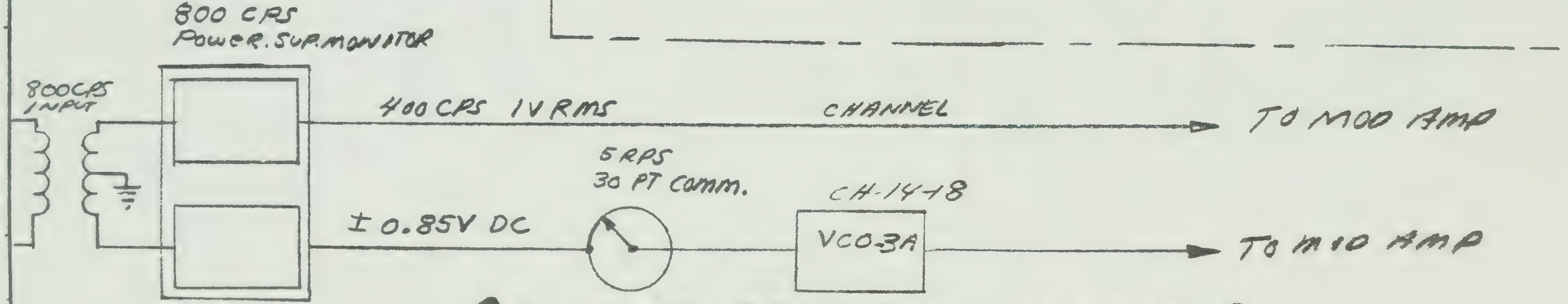
T/M INFLIGHT CALIBRATE CHECK OUT KLUGE

56-3-h

VOLTAGE MONITOR

INPUT*	output
100V	+0.85V
115V	0V
130V	-0.85V





800 CPS FREQ & VOLT MONITOR

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LOCKHEED AIRCRAFT CORPORATION
MISSILE SYSTEMS DIVISION
XN NAVY TRAINING SCHOOL

Contract NOp-834
Course Notes SC-3-i

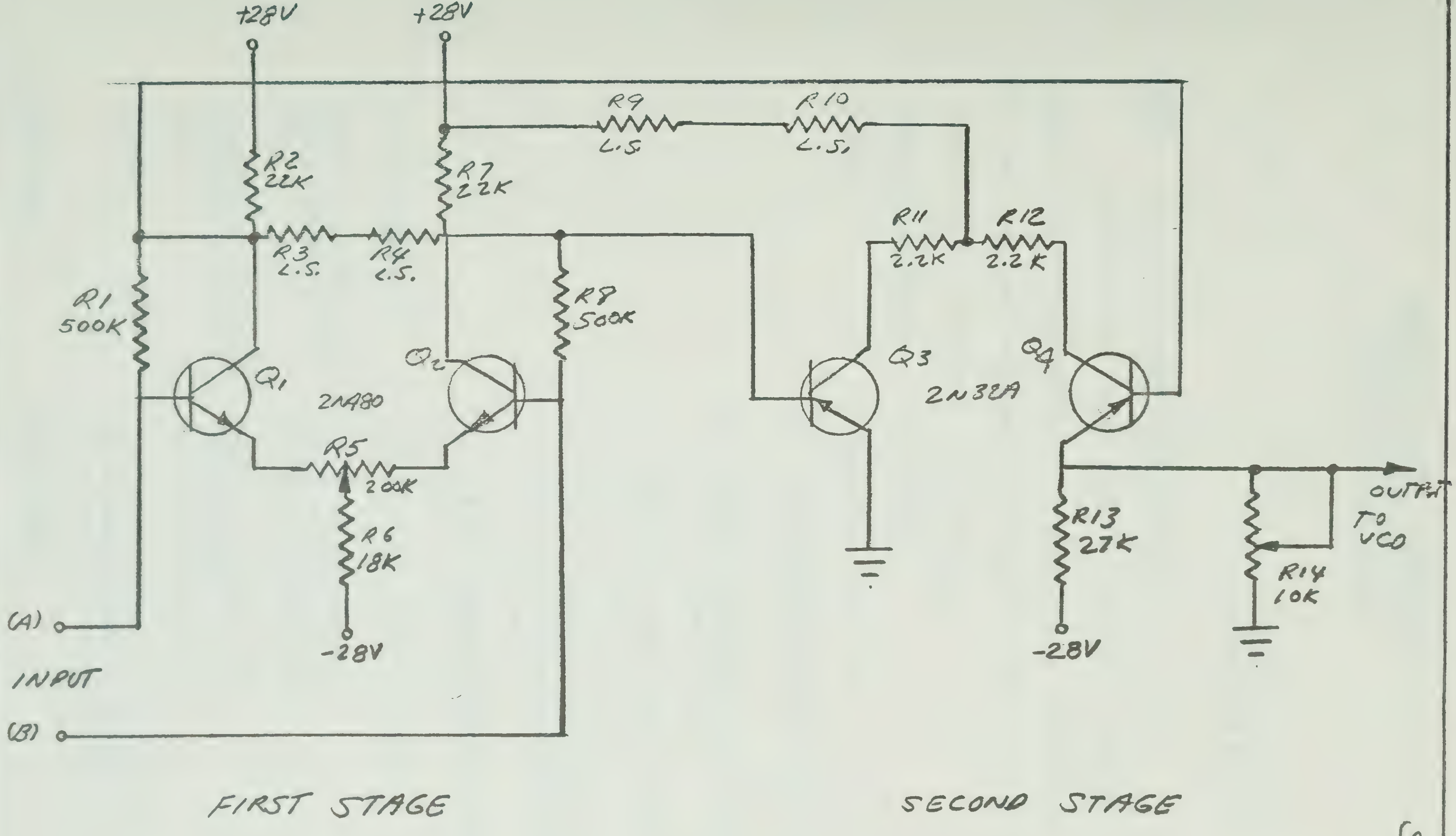
TOPIC: D. C. Amplifier Characteristics

This unit will be capable of amplifying a low input voltage to the level required for operation of a VCO. The tolerances to which it must conform are those listed below:

Power Requirement:	28 VDC ±1%; 5.0 ma max. 28 VDC ±1%; 3.0 ma max (ungrounded)
Input Requirement:	0 to 50 millivolts from thermocouple or strain gage transducer.
Input Impedance:	1K ohms minimum
Output Impedance:	5K ohms maximum
Signal Output:	Single Ended
Frequency Response:	Flat within 3 db from DC to 1.5 KC
Gain:	45 min - 55 max
Gain Stability:	± 2.0%
Drift Stability:	± 2.0%
Linearity:	± 1.0% max deviation from best straight line
Output voltage with open input:	Maximum output shall not exceed 2.0 volts RMS

Reference Well: This component serves to measure the temperature at the thermocouple lead junction which forms the transition from thermocouple metal conductors to copper conductors. This permits calculation of the actual hot junction temperature.

Power Requirements:	Resistance Thermometer excitation Current
Accuracy:	±3% of temperature measurement. (Permits calculation of hot junction temperature to better than 1%)



D.C. Amplifier
FIG. 1

SC-3-j

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LOCKHEED AIRCRAFT CORPORATION
MISSILE SYSTEMS DIVISION
XN NAVY TRAINING SCHOOL

Contract NOp-834
Course Notes SC-3-j

TOPIC: DC to DC Converter Characteristics

This component furnishes all operating voltages for system operation and incorporates regulation as dictated by individual component requirements. Operation shall conform to the following minimum standards of performance:

Power Requirements:	28 VDC ±10%
Output:	The outputs shall be: a. +28 VDC preset within ±2% at 500 ma. load. b. 12 VDC ungrounded-preset within 2% at 3 ampere load. c. +350 VDC preset within ±2% at 110 ma. load. d. Six (6) separate ungrounded outputs of 30 VDC preset within ±2% at 5 ma. load. e. 30 VDC ungrounded-preset within ±2% at 40 ma. load.
Output Impedance:	1.0 ohm max. from 0 to 50 KC
Regulation with 10% input voltage variation	±0.5% on outputs a & b ±1.5% on output c ±0.5% on outputs d & e
Regulation with load Variation from 75 to 100%	±1.0% on outputs a & b ±3% on output c ±1.0% on outputs d & e
Ripple:	5 mv peak-to-peak on outputs a & b 50 mv peak-to-peak on output c 20 mv peak-to-peak on outputs d & e
Efficiency:	60% minimum

Overload:	No degradation in performance after 100% overload for 30 sec.
Short circuit:	Capable of withstanding an output short circuit of one second at ambient temperature with full recovery in 500 milli-seconds, when operated from battery or magnetic supply.

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LOCKHEED AIRCRAFT CORPORATION
MISSILE SYSTEMS DIVISION
XN NAVY TRAINING SCHOOL

Contract NOp-834
Course Notes SC-3-k

TOPIC: Radio Frequency Monitor Characteristics:

The RF Monitor samples the RF power being transmitted to the antenna complex for a relative indication of any change in transmitted power. This component shall conform to the following performance requirements:

Power Requirement:	None
Signal Output:	0.5 VDC minimum for a power transfer of 5 watts in the transmission line
Standing Wave Ratio:	5% maximum increase in transmission line SWR to be caused by insertion of RF monitor.

TOPIC: Squib Monitor Characteristics:

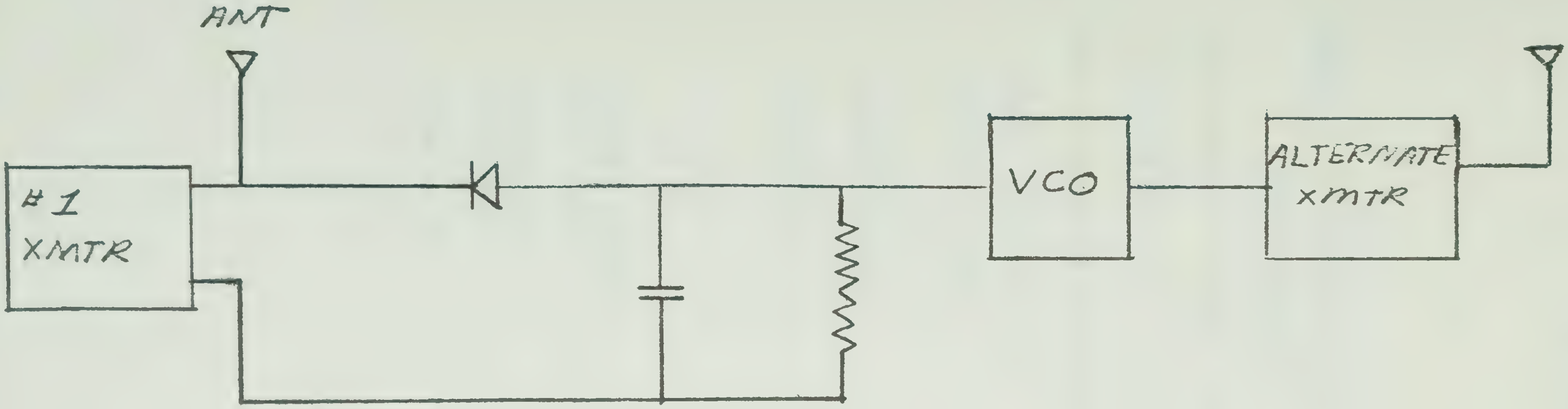
The squib current monitor serves to indicate squib firing by monitoring occurrence and nominal amplitude of the current pulse which fires the squib.

Power Requirement:	None
Input Requirement:	28 VDC ± 10% @ 20 amperes excitation current
Polarity:	Positive excursion of leading waveform at filter input with positive input voltage relative to ground.
Output Load:	25 K ohms ± 5%
Output Pulse Amplitude:	50 volt peak ± 30 volts
Output Pulse Duration:	Output pulse shall drop 63% from peak value within 0.5 ± 0.2 milliseconds.

TOPIC: Transducer Switch Characteristics:

The purpose of this component is to transfer the source of data information from one arbitrary group of transducers to a second group of transducers by command or specific function occurrence.

Coil Requirement:	28 VDC ±10% @ 30 ma maximum per relay
Contact Resistance:	0.1 ohm maximum over repeated actuations at 10 ma current through contacts.



RF-MONITOR

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MISSILE SYSTEMS DIVISION
XN NAVY TRAINING SCHOOL

Contract NOp-834
Course Notes SC-3-1

TOPIC: Commutator Characteristics

This component is a motor driven sampling switch which serves to sample numerous data on a time sharing basis for application to a single subcarrier channel. The unit shall conform to the following performance requirements.

Power Requirements:	28VDC ± 10% @ 13 watts max.
Sampling Rate:	±10% of specified sampling rate at rated voltage
Pole Phasing:	1 degree of wiper rotation
Duty cycle:	20% maximum shorting time 80% minimum non-shorting time (for make-before break) 60% minimum "on" time on break-before-make type
Average Contact Resistance:	1 ohm maximum
Peak Contact Resistance:	3 ohms maximum peak-to-peak as referenced to average
Service Life:	200 hours minimum service free life without degradation of performance as specified.

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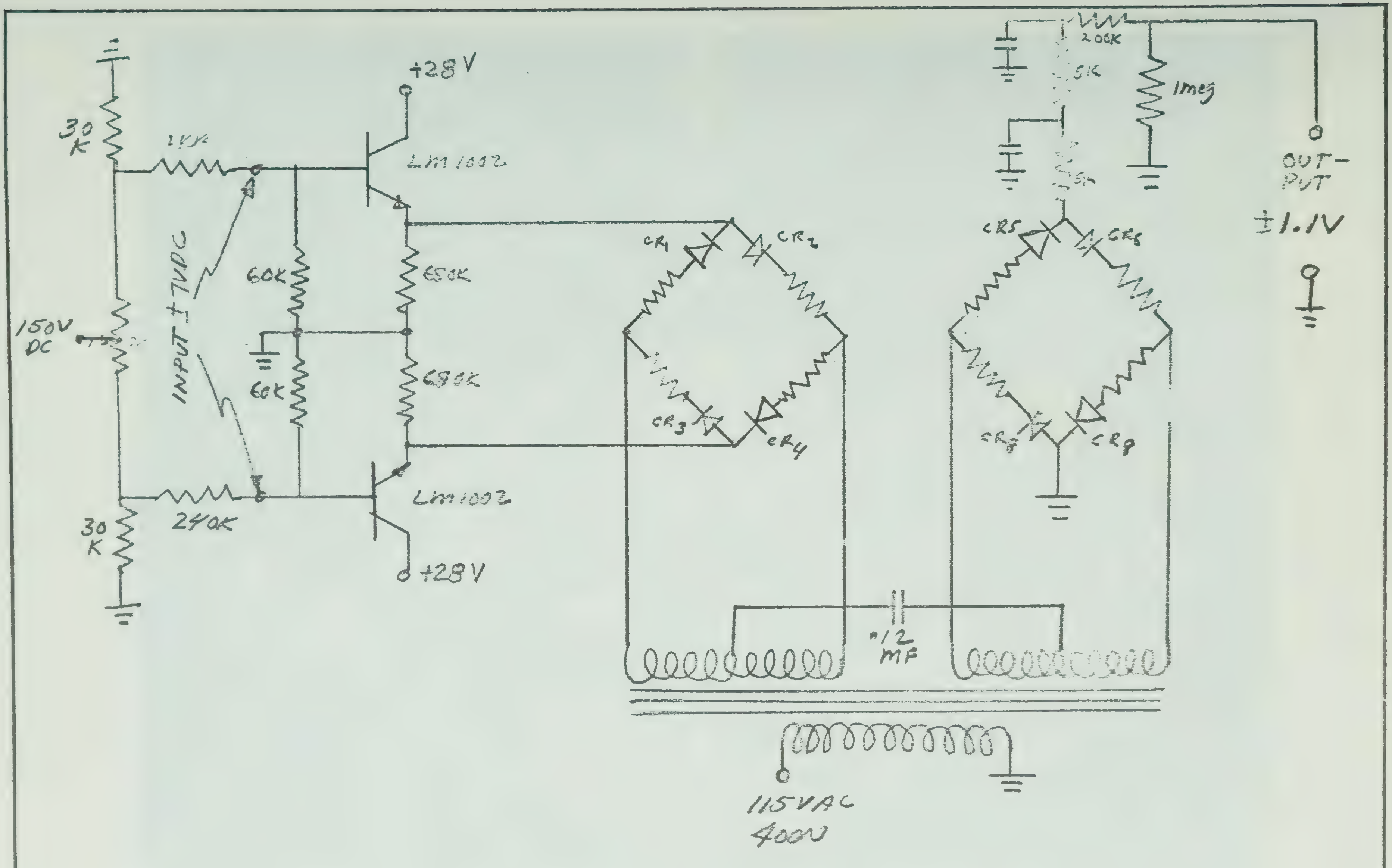
MISSILE SYSTEMS DIVISION
XN NAVY TRAINING SCHOOL

Contract NOp-834
Course Notes SC-3-m

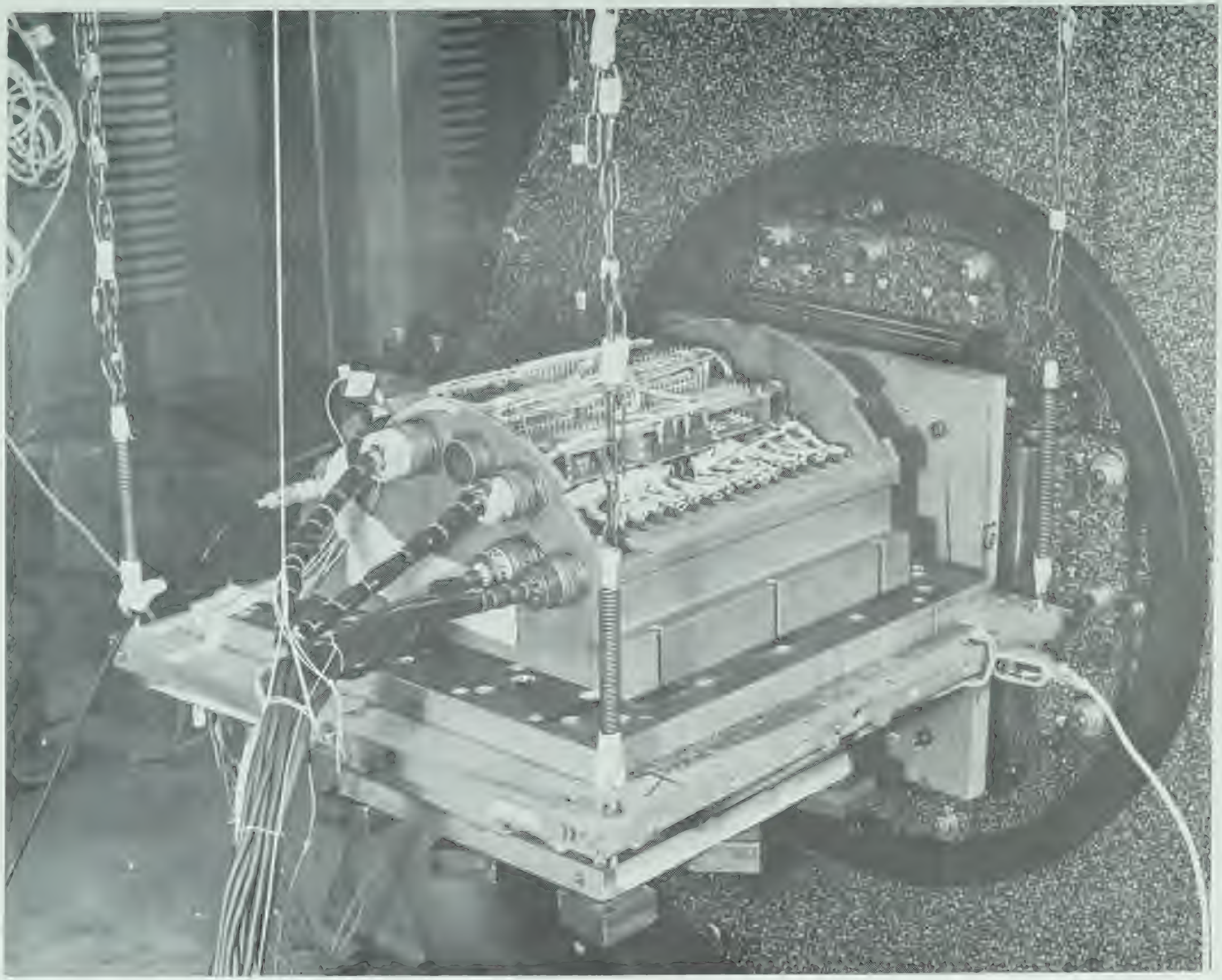
TOPIC: Valve Current Monitor Characteristics

This component serves to measure the differential output voltage of a hydraulic valve at a high voltage relative to ground.

Power Requirement:	+28 VDC ± 10% 115 VAC @ 400 CPS
Input Signal:	0 to ± 7 VDC at a maximum of 0.5 ma
Input Impedance:	15 K ohms minimum
Output Load:	250 K ohms
Output Voltage:	± 1.1 VDC + 0.2 or - 0.3V
Linearity:	1.5% of full scale output
Zero Offset:	0 ± 0.2 VDC with input shorted
Output Ripple:	1% maximum of full scale output

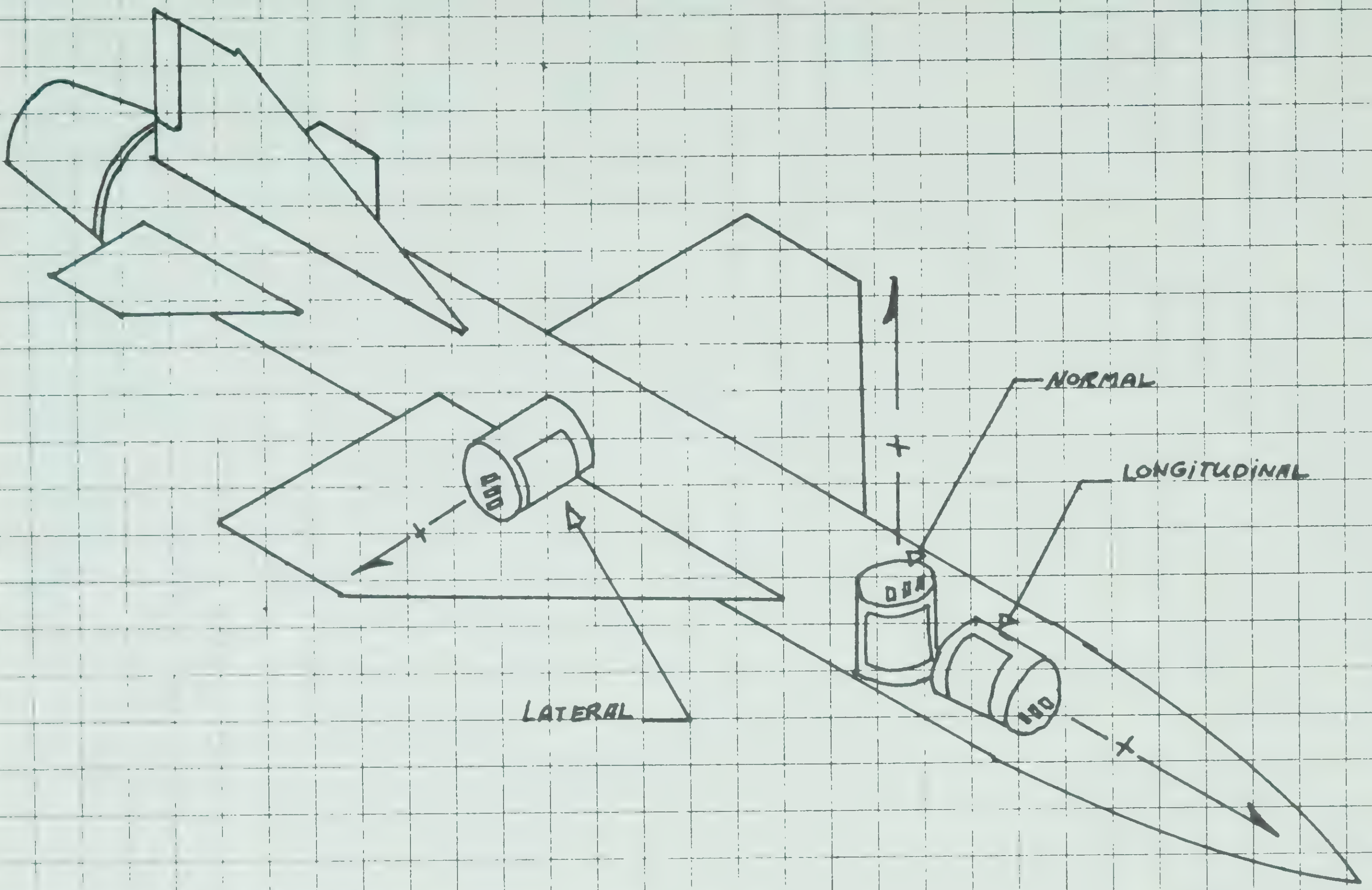


VALVE CURRENT MONITOR

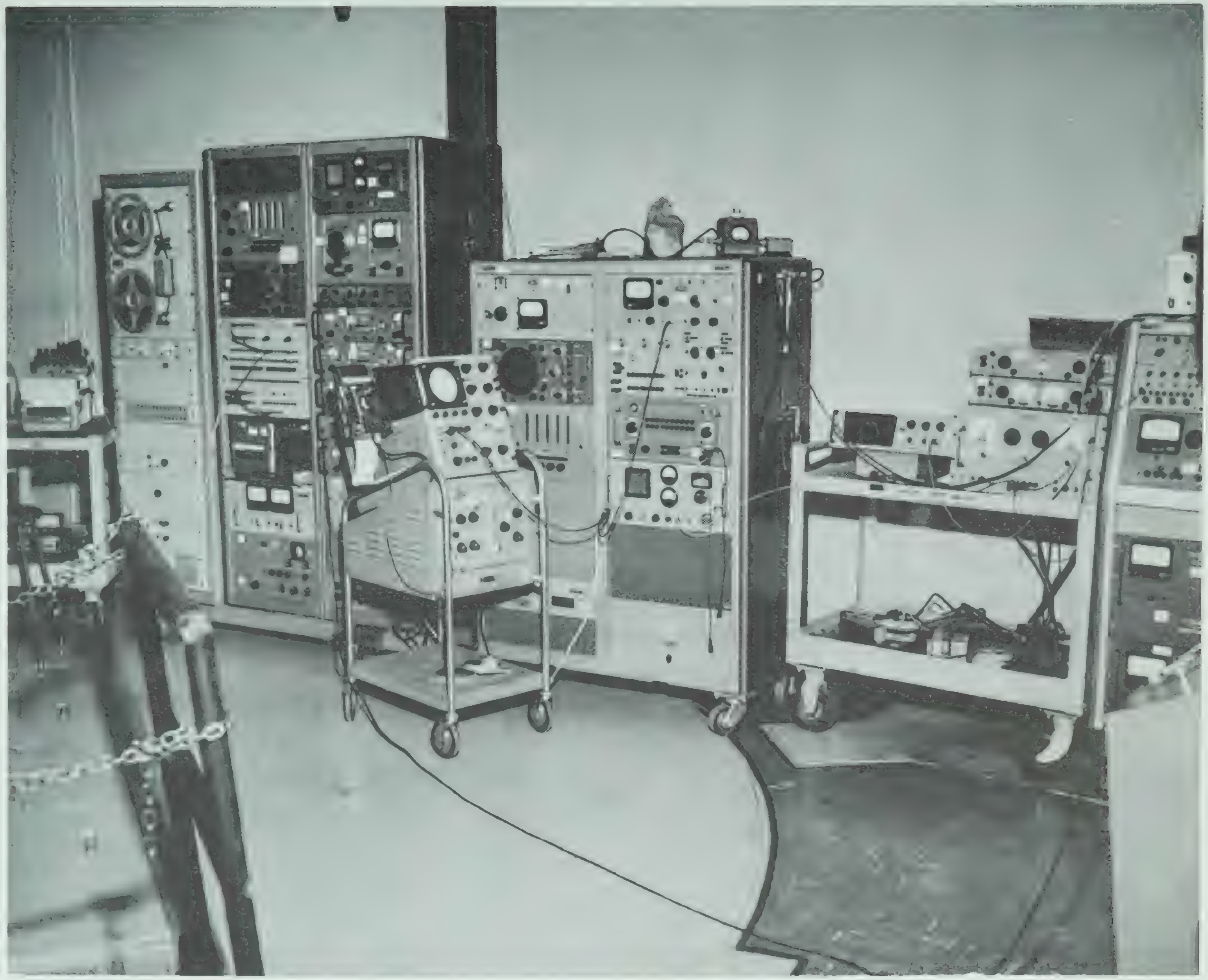


Link #1 or #2 T/M Package

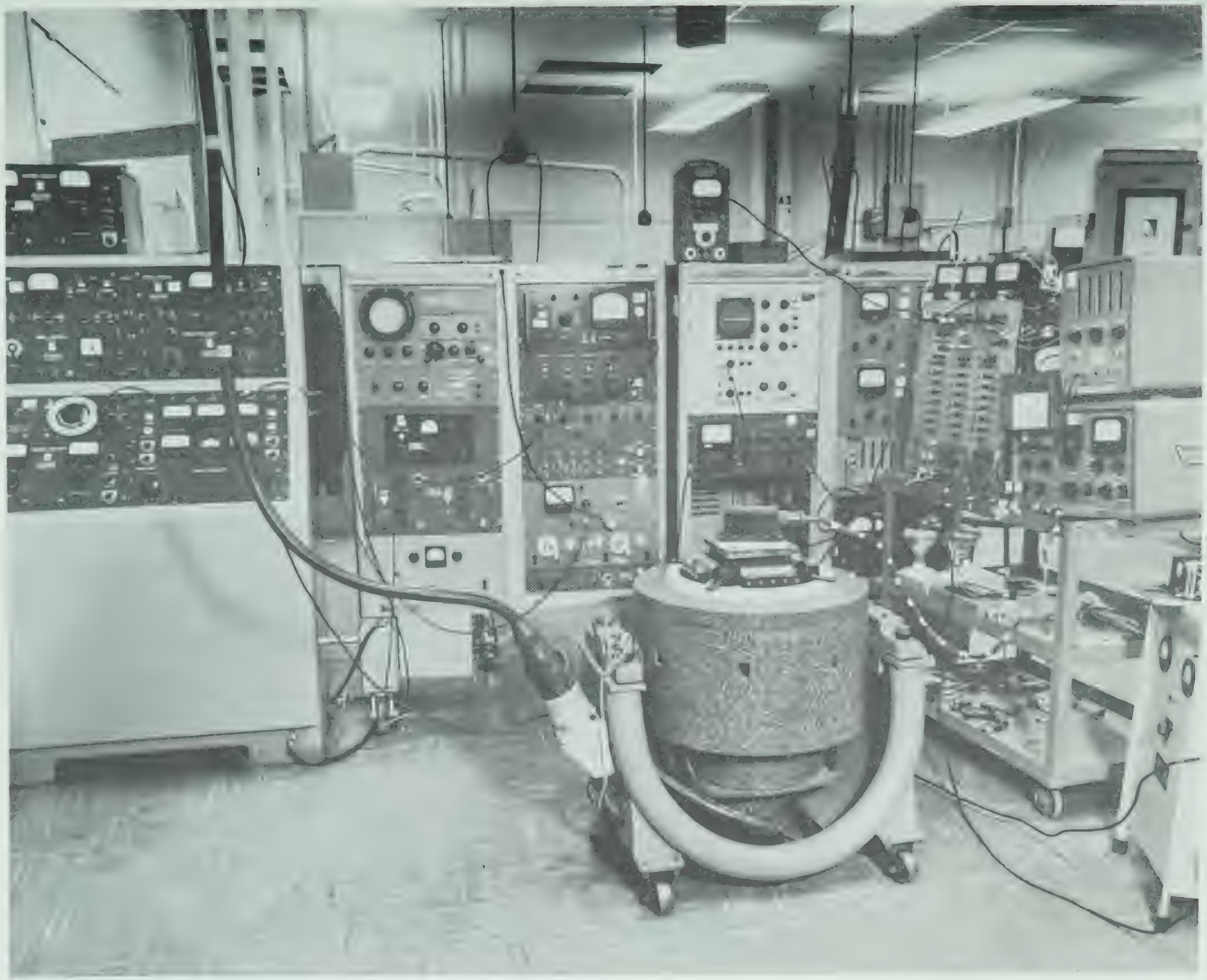
	NAME	DATE	LOCKHEED AIRCRAFT CORPORATION MISSILE SYSTEMS DIVISION	Page	TEMP.	PERM.
Prepared						
Checked			TITLE	Model		
Approved				Report No.		



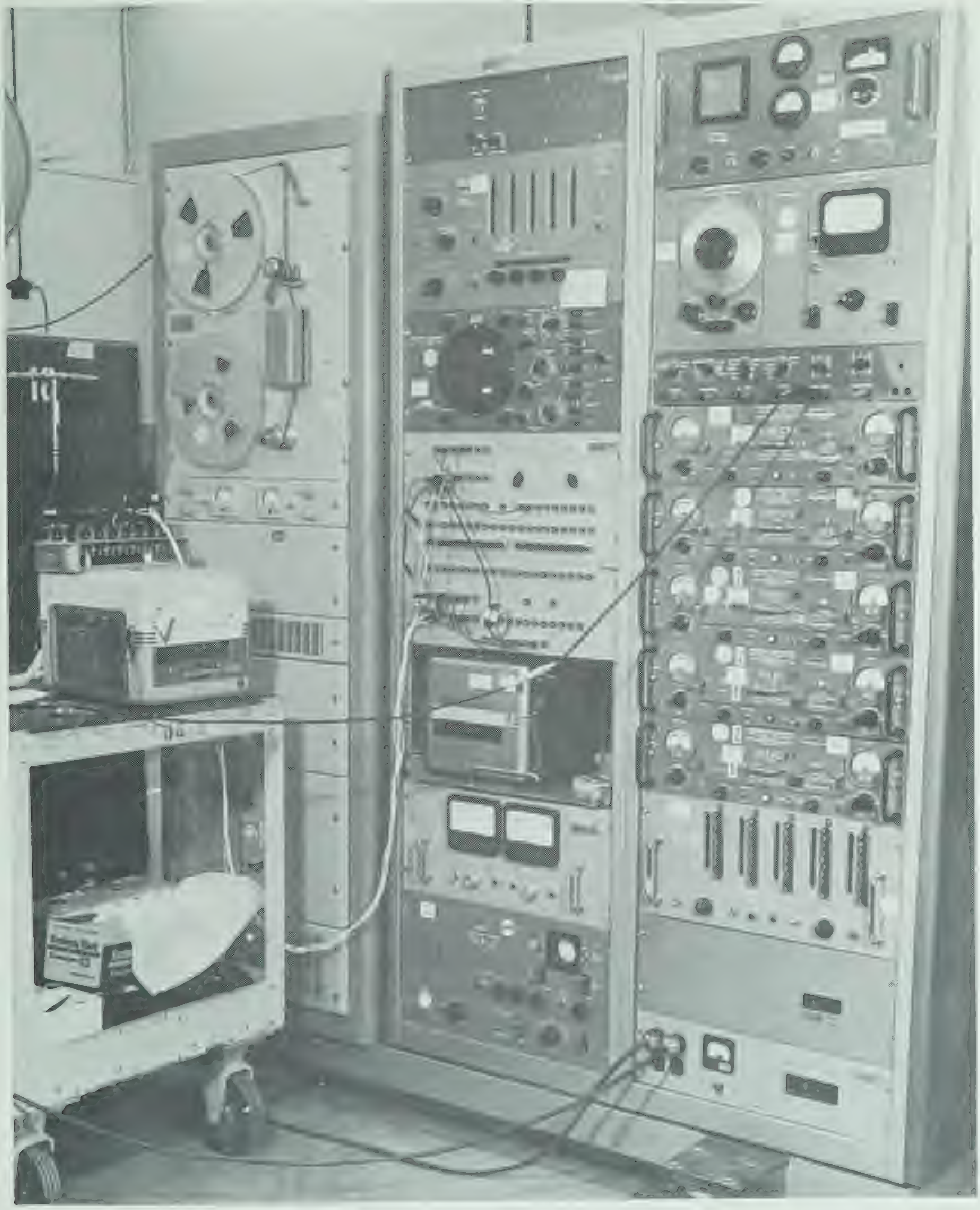
STANDARD ACCELEROMETER MOUNTING



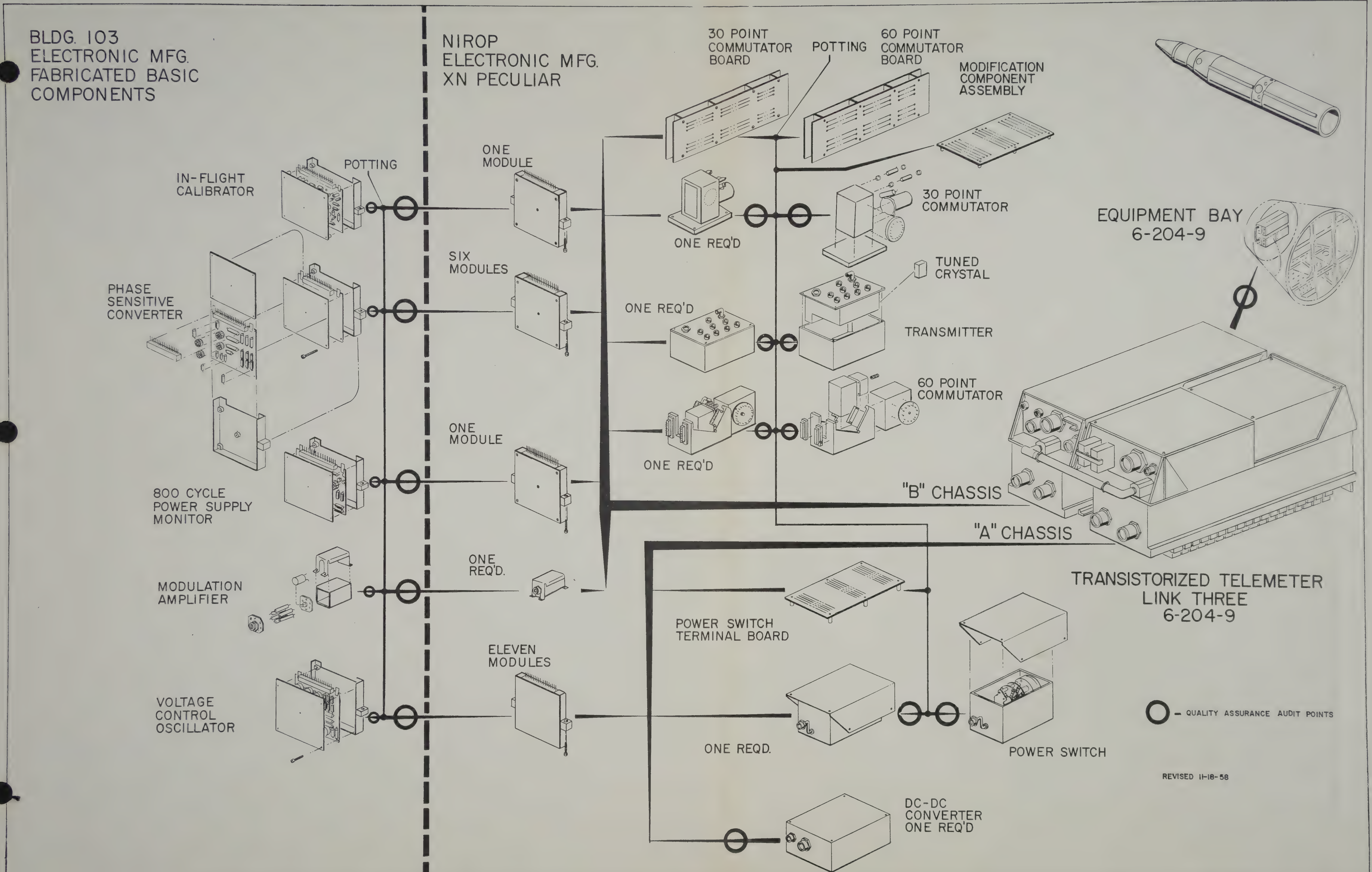
T/M Package Power Supplies
and Receiving Equipment

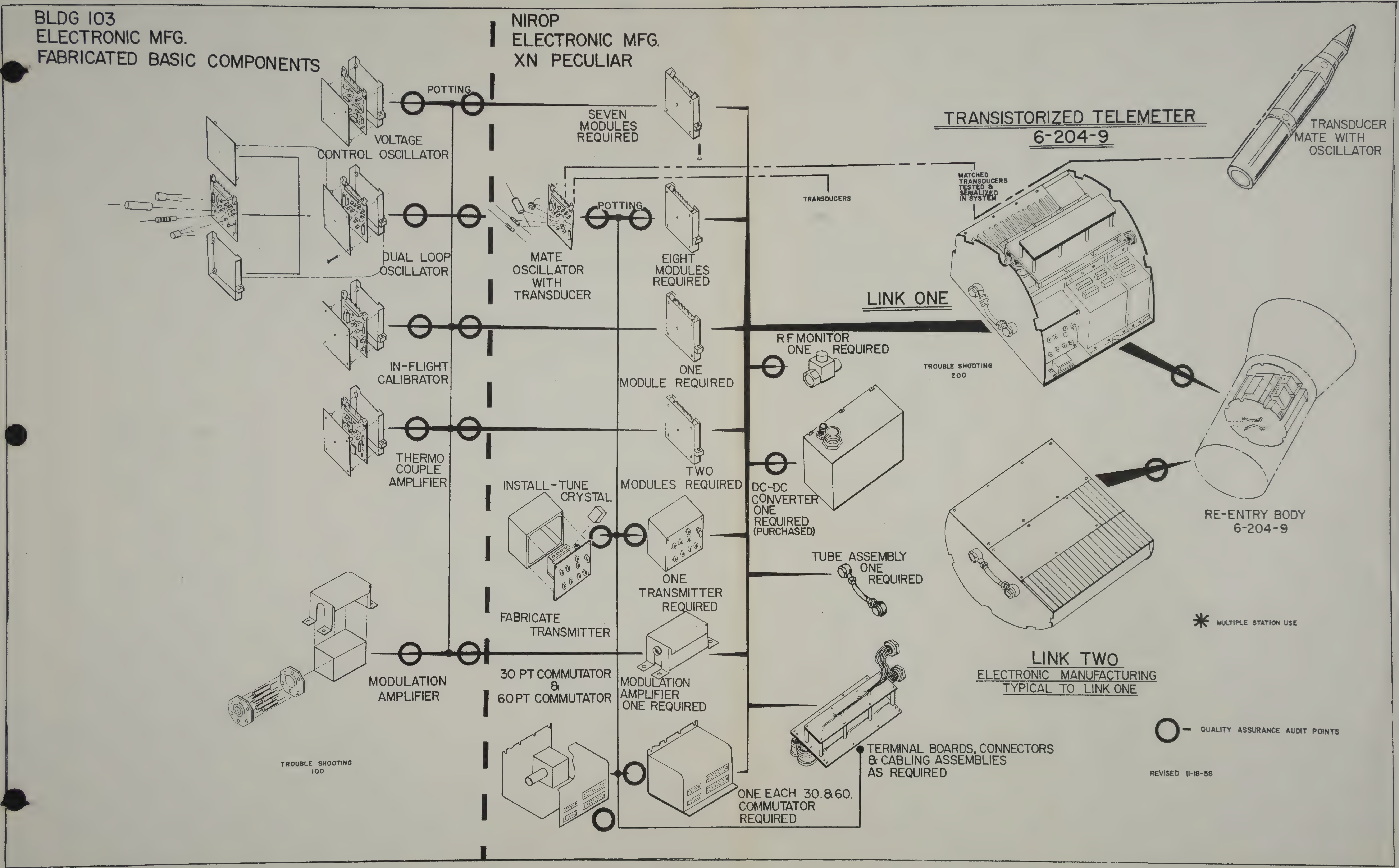


Environmental Testing Equipment

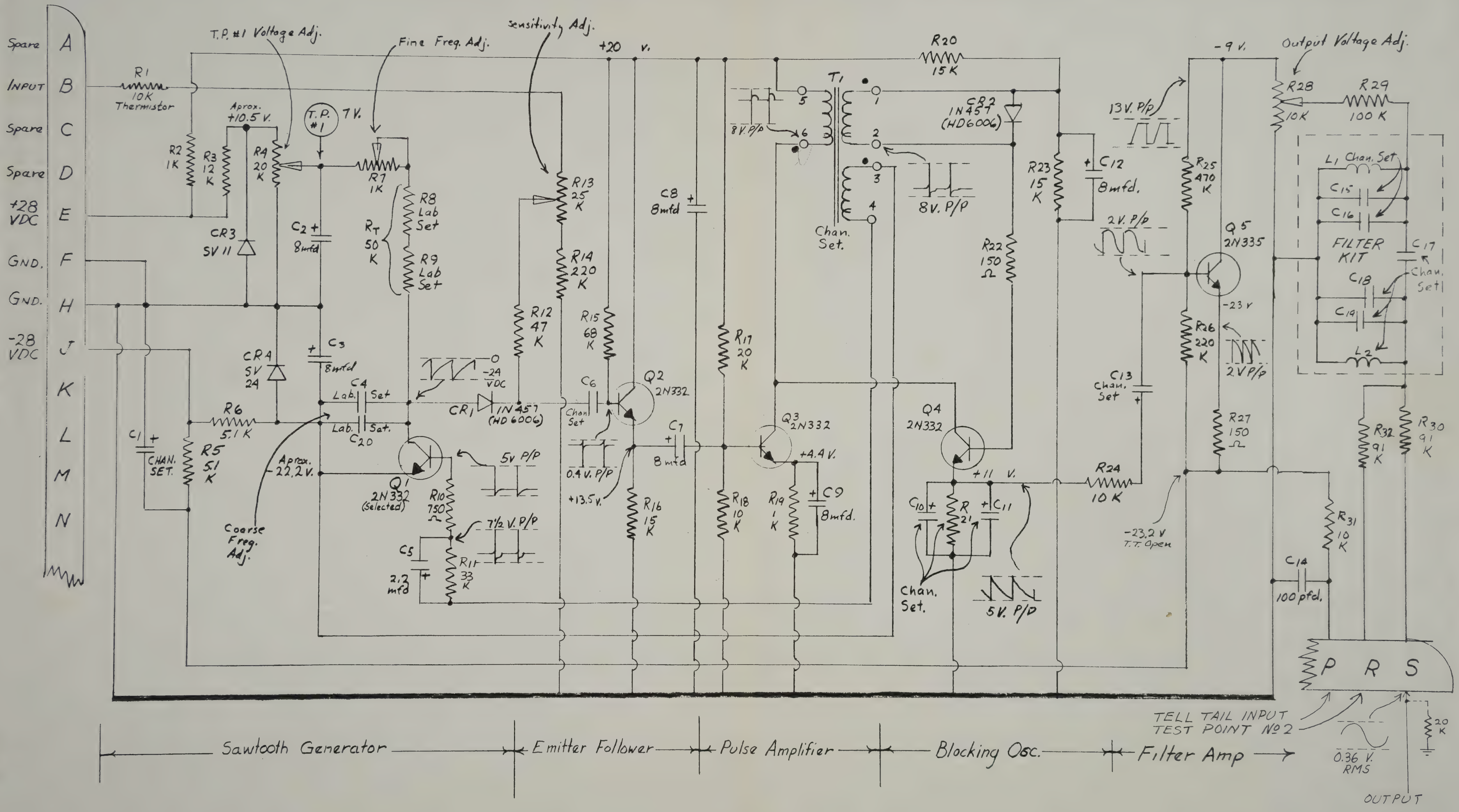


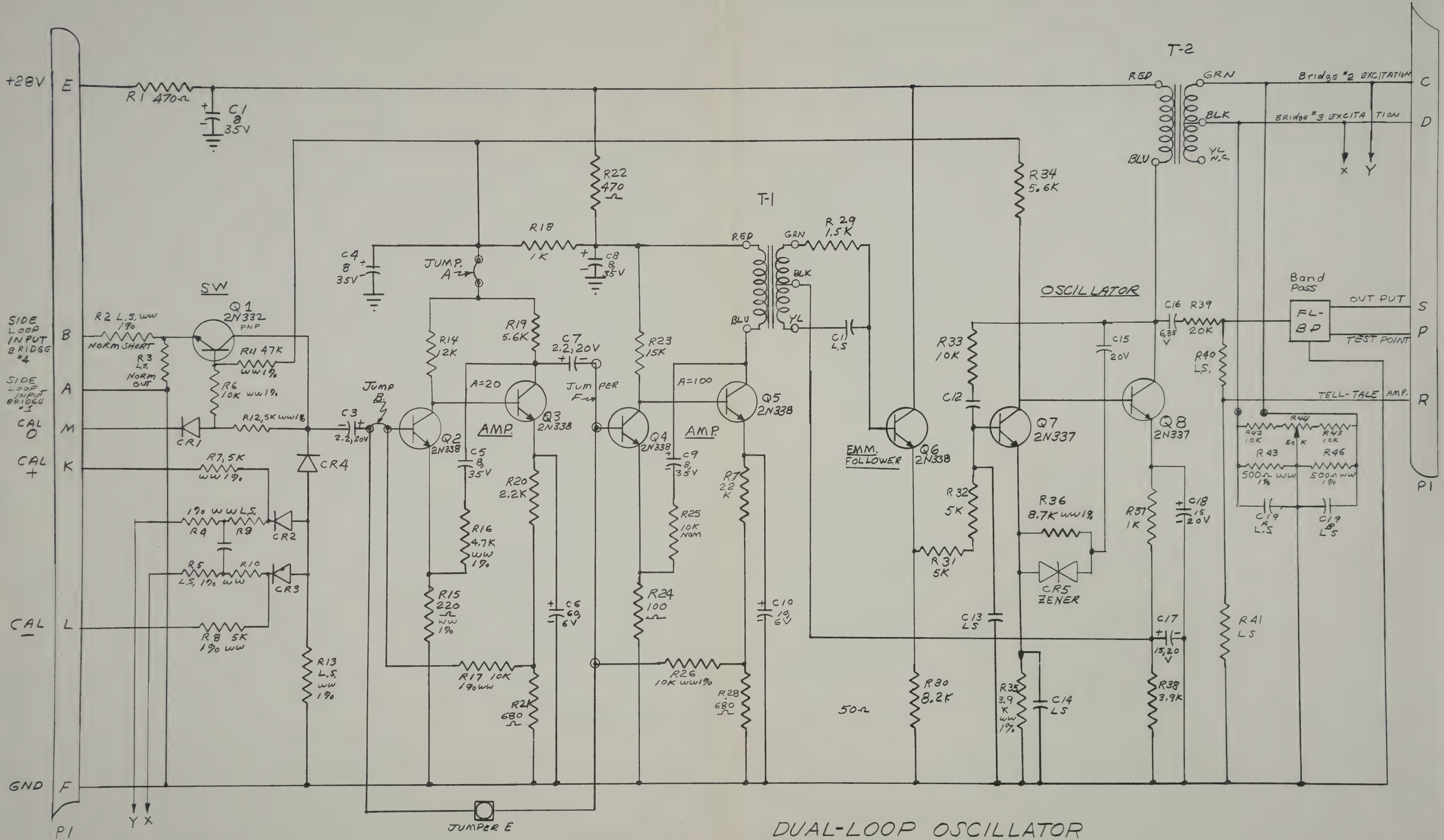
Tape Recorder and Sub-Carrier
Discriminators


BLDG. 103
ELECTRONIC MFG.
FABRICATED BASIC
COMPONENTS
NIROP
ELECTRONIC MFG.
XN PECULIAR
30 POINT COMMUTATOR BOARD
POTTING
60 POINT COMMUTATOR BOARD
MODIFICATION COMPONENT ASSEMBLY
POTTING
IN-FLIGHT CALIBRATOR
ONE MODULE
30 POINT COMMUTATOR
ONE REQ'D
EQUIPMENT BAY
6-204-9
SIX MODULES
PHASE SENSITIVE CONVERTER
ONE REQ'D
TUNED CRYSTAL
TRANSMITTER
60 POINT COMMUTATOR
ONE MODULE
ONE REQ'D
800 CYCLE POWER SUPPLY MONITOR
"B" CHASSIS
"A" CHASSIS
ONE REQ'D.
MODULATION AMPLIFIER
POWER SWITCH TERMINAL BOARD
TRANSISTORIZED TELEMETER
LINK THREE
6-204-9
ELEVEN MODULES
VOLTAGE CONTROL OSCILLATOR
ONE REQD.
POWER SWITCH
○ - QUALITY ASSURANCE AUDIT POINTS
REVISED 11-18-58
DC-DC CONVERTER
ONE REQ'D


BLDG 103
ELECTRONIC MFG.
FABRICATED BASIC COMPONENTS
NIROP
ELECTRONIC MFG.
XN PECULIAR
POTTING
SEVEN MODULES REQUIRED
VOLTAGE CONTROL OSCILLATOR
TRANSISTORIZED TELEMETER
6-204-9
TRANSDUCER MATE WITH OSCILLATOR
MATCHED TRANSDUCERS TESTED & SERIALIZED IN SYSTEM
TRANSDUCERS
POTTING
DUAL LOOP OSCILLATOR
MATE OSCILLATOR WITH TRANSDUCER
EIGHT MODULES REQUIRED
LINK ONE
RF MONITOR ONE REQUIRED
IN-FLIGHT CALIBRATOR
ONE MODULE REQUIRED
TROUBLE SHOOTING 200
THERMO COUPLE AMPLIFIER
TWO MODULES REQUIRED
DC-DC CONVERTER ONE REQUIRED (PURCHASED)
INSTALL-TUNE CRYSTAL
RE-ENTRY BODY
6-204-9
TUBE ASSEMBLY ONE REQUIRED
ONE TRANSMITTER REQUIRED
FABRICATE TRANSMITTER
MULTIPLE STATION USE
MODULATION AMPLIFIER
30 PT COMMUTATOR & 60 PT COMMUTATOR
MODULATION AMPLIFIER ONE REQUIRED
LINK TWO
ELECTRONIC MANUFACTURING TYPICAL TO LINK ONE
QUALITY ASSURANCE AUDIT POINTS
TROUBLE SHOOTING 100
TERMINAL BOARDS, CONNECTORS & CABLING ASSEMBLIES AS REQUIRED
REVISED 11-18-58
ONE EACH 30.& 60. COMMUTATOR REQUIRED

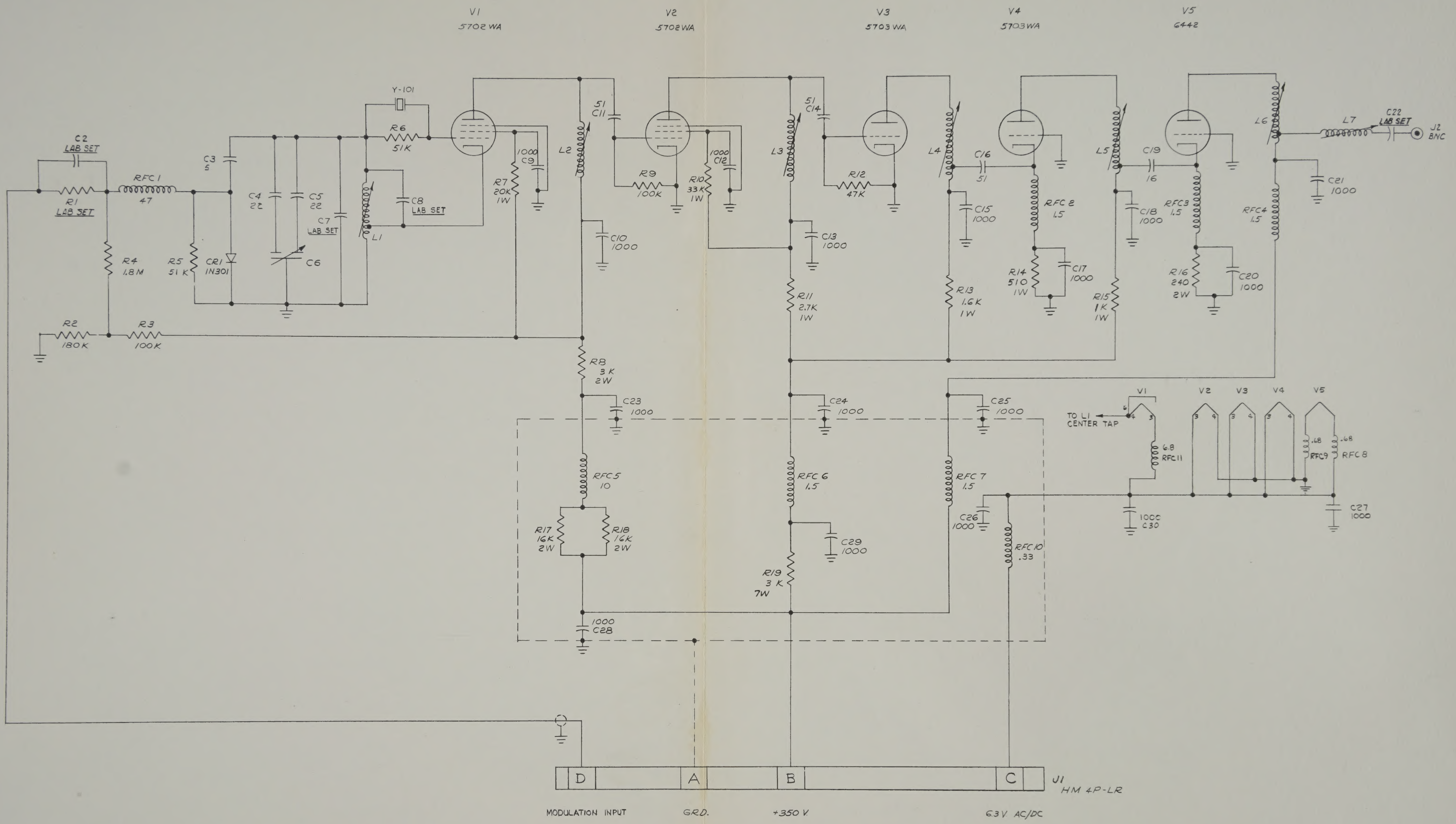
SCHEMATIC DIAGRAM TR-VCO-3A 8-6-58 M.T. WILLIAMS (Revised 2-6-59 B. Mendenhall)





DUAL-LOOP OSCILLATOR

POLARIS TRAINING — Stone



V1
5702 WA
V2
5702 WA
V3
5703 WA
V4
5703 WA
V5
6442
Y-101
C2
LAB SET
R1
LAB SET
RFC1
47
C3
5
R6
51K
C4
22
C5
22
C7
LAB SET
C8
LAB SET
L1
R4
1.8M
R5
51 K
CR1
1N301
C6
R2
180K
R3
100K
R7
20K
1W
1000
C9
51
C11
L2
C10
1000
R9
100K
R10
33K
1W
1000
C12
51
C14
L3
C13
1000
R11
2.7K
1W
R12
47K
L4
C16
51
C15
1000
RFC 2
1.5
R14
510
1W
C17
1000
R13
1.6K
1W
L5
C19
16
C18
1000
RFC3
1.5
R16
240
2W
C20
1000
R15
1K
1W
L6
L7
C22
LAB SET
J2
BNC
C21
1000
RFC4
1.5
R8
3 K
2W
C23
1000
C24
1000
C25
1000
RFC5
10
R17
16K
2W
R18
16K
2W
RFC 6
1.5
C29
1000
R19
3 K
7W
RFC 7
1.5
C26
1000
RFC10
.33
1000
C28
V1
TO L1
CENTER TAP
6.8
RFC11
1000
C30
V2
V3
V4
V5
.68
RFC9
.68
RFC8
C27
1000
D
A
B
C
J1
HM 4P-LR
MODULATION INPUT
GRD.
+350 V
6.3V AC/DC

LOOK TO LOCKHEED FOR LEADERSHIP